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FOREIGN TECHNOLOGY DIVISION

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BASES OF RADIO DIRECTION FINDING

by

I. S. Kukes, M. Ye. Starik

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U. S. BOARD ON GEOGRAPHIC NAMES TRANSLITERATION SYSTEM

Block	Italic	Transliteration	Block	Italic	Transliteration
А а	*А а*	A, a	Р р	*Р р*	R, r
Б б	*Б б*	B, b	С с	*С с*	S, s
В в	*В в*	V, v	Т т	*Т т*	T, t
Г г	*Г г*	G, g	У у	*У у*	U, u
Д д	*Д д*	D, d	Ф ф	*Ф ф*	F, f
Е е	*Е е*	Ye, ye; E, e*	Х х	*Х х*	Kh, kh
Ж ж	*Ж ж*	Zh, zh	Ц ц	*Ц ц*	Ts, ts
З з	*З з*	Z, z	Ч ч	*Ч ч*	Ch, ch
И и	*И и*	I, i	Ш ш	*Ш ш*	Sh, sh
Й й	*Й й*	Y, y	Щ щ	*Щ щ*	Shch, shch
К к	*К к*	K, k	Ъ ъ	*Ъ ъ*	"
Л л	*Л л*	L, l	Ы ы	*Ы ы*	Y, y
М м	*М м*	M, m	Ь ь	*Ь ь*	'
Н н	*Н н*	N, n	Э э	*Э э*	E, e
О о	*О о*	O, o	Ю ю	*Ю ю*	Yu, yu
П п	*П п*	P, p	Я я	*Я я*	Ya, ya

*ye initially, after vowels, and after ъ, ь; e elsewhere.
When written as ё in Russian, transliterate as yё or ё.

RUSSIAN AND ENGLISH TRIGONOMETRIC FUNCTIONS

Russian	English	Russian	English	Russian	English
sin	sin	sh	sinh	arc sh	$\sinh^{-1}$
cos	cos	ch	cosh	arc ch	$\cosh^{-1}$
tg	tan	th	tanh	arc th	$\tanh^{-1}$
ctg	cot	cth	coth	arc cth	$\coth^{-1}$
sec	sec	sch	sech	arc sch	sech^{-1}
cosec	csc	csch	csch	arc csch	csch^{-1}

Russian	English
rot	curl
lg	log

Page 1.

BASES OF RADIO DIRECTION FINDING.

I. S. Kukes, M. Ye. Starik.

Page 2.

In the book are presented the principles and the methods of radio traffic, are described the different systems of radio direction finders, are given the methods of the calculation of radio direction finders and their antenna systems. Are examined the errors of radio direction finders and special feature/peculiarity of installation, checking and use of radio direction finders under varied conditions.

The book is intended for the students of schools of higher education as textbook to the course of radio direction finders, for students and radio engineers as management/manual on design and calculation of the different systems of radio direction finders, and also for the operating personnel, which operates direction-finding equipment and of its realizing organization use.

The book will be useful for the wide circle of the radio specialists, connected with that which was directed by radio reception.

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PREFACE.

Radio direction finders widely are used in air and marine transport for the solution of navigational problems (position finding of movable object, flight toward airport, motion to the ship, that suffers calamity, and so forth); they are applied also for other target/purposes (research on the questions of radiowave propagation, observation of space vehicles and so forth).

Known direction-finding methods continuously are improved, increasingly more deeply are develop/processed the paths of an increase in the accuracy and sensitivity, are found new direction-finding methods, is improved theory, are expanded the frequency band and the field of application of direction finders. All the enumerated questions are illuminated in periodic technical literature; however, until now, are not systematized.

In the book is presented the general theory of direction finding, are given the procedures of calculation of direction finder and its cell/elements, are analyzed the errors of direction finding and way of their elimination.

The authors hope that the proposed book will be useful not only for persons, occupied with development and the use of direction finders, but also to the wide circle of the radio specialists whose activity is connected with the directed radio reception.

The authors consider it their debt to express deep appreciation to Candidate of Technical Sciences V. K. Mezin for writing §8.3-8.6 and the survey of the manuscript, to Candidates of Technical Sciences L. Sh. Natadze and to V. N. Ivanov for a series of observations and indications, made during the review of the book.

Page 4. No typing.

Page 5.

Chapter 1.

PROBLEMS OF RADIO TRAFFIC.

Radio traffic was used for the first time and in essence was developed as means of the marine, and then also air navigation, for which the most important question is the determination of the position of moving object (ship, aircraft). For determining the position of any of object, it is necessary to determine angles with certain reference direction of straight lines [1], that connect this object with the points whose coordinates are accurately known.

FOOTNOTE [1] since the surface of terrestial globe plane, precisely to say about the geodetic lines, which connect the given points with this surface. For greater detail, see chapter 12. ENDFOOTNOTE.

If in point x is located the object whose position is determined, and the coordinates of point A are known, then direction from point x into point A, determined by angle α between the direct/straight, connecting point x and A (Fig. by 1.1), and certain reference directions, he is called bearing. As the reference direction from which are counted off all angles, is accepted usually the direction of true (geographical) meridian at the particular point. In that case bearing he is called true.

For determining the position of any of object, it is necessary to determine from the point of the position of object x the bearings of two points A and B, for example α and β (see Fig. 1.1). After finding these bearings, construct on map/chart straight lines AE and BD, forming at points A and B with direction north - south angles $\alpha' = \alpha - 180°$ and $\beta' = \beta + 180°$ respectively.

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Intersection of the straight lines AE and BD gives direct position of point x. The directions, determined by angles α' and β', they are called reciprocal bearings.

Bearings can be determined by visual and optical methods. These methods have two essential deficiency/lacks: the small range, limited by line-of-sight ranges, and the impossibility of their use under conditions of poor visibility, i.e., when the precision determination of the position of moving object most of all is necessary.

It is natural that for this purpose were used the instruments, determining bearings by the means of radio, the radio direction finders, possessing the considerably larger range and the possibility of work in fog and under other conditions of poor visibility.

The bearing, determined with the aid of radio equipment, he is called radio bearing. The process of determining the radio bearing he is called radio traffic, and the branch of radio engineering, which studies all questions, connected with radio traffic, by radio direction finding.

In the preceding/previous example it was assumed that the determination of directions α and β is conducted on the moving object itself. This method of the direction finding when direction finder is located on ship or aircraft, he is called its own direction finding.

With direction finding it is possible to utilize a transmission of any radio station whose position is accurately known. For the purpose of the provision for a possibility of direction finding at any time on the earth/ground, are establish/installed the special transmitting radio stations, called the nondirectional radio beacons.

Besides the described method is applied still another method of direction finding, which lies in the fact that with the aid of the direction finders, arrange/located on the earth/ground, are determined the angles α' and β', which are reciprocal bearings: $\alpha' = \alpha - 180°$; $\beta' = \beta + 180°$.

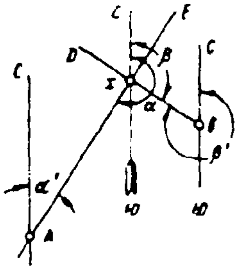

Fig. 11. Determination of position from two bearings.

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This method he is called strange direction finding. For the possibility of its application/use, it is necessary on aircraft (or ship) to have the radio-transmitter station for the call of terrestrial radio direction finding station and then for emission/radiation during direction finding. The obtained bearings terrestrial station radioed aboard aircraft or ship.

A main deficiency/lack in this method lies in the fact that the services of direction-finding station at the given instant can use only one object, while the nondirectional radio beacon can be oriented by which conveniently number of aircraft or ships, equipped with direction finders.

The advantage of the method of strange direction finding consists in the fact that it can use the ships and aircraft, which do not have special equipment (radio direction finder), but equipped only with the normal receiveting-transmit radio station.

Besides the determination of two directions for determining position in navigation, has value and definition of one direction along which must follow the aircraft or ship. This target/purpose serve, first of all, compass, also, at small distances - the so-called double beacons. Here also with large success can be used radio direction finder.

In some special cases the role of radio direction finder can be especially important, for example, the direction of ship in aid to other, signalling of calamity. If the location of the latter is known insufficiently accurately, the only method to rapidly achieve it is floating in the direction, indicated by direction finder. Large conveniences direction finder represents also when conducting of the caravan of law courts by ice-breaker, etc.

In the air fleet the radio direction finders widely are applied for the solution of the problems of air navigation, in the traffic control service of the provision for aircraft guidance, for the identification of aircraft in airport zone and in other cases.

Besides navigation the radio direction finders find a use in military science as means of determining the position of radio stations and, therefore, troop formations of enemy.

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For this purpose, the direction finders widely and successfully were applied even in world war 1914-1918.

Finally, radio traffic is very essential method for research on a number of the physical problems, connected with radio engineering, mainly the questions of the propagation of the electromagnetic waves of different range, in different time of days and year, distribution of atomospheric discharges, etc.

The important application/use of radio direction finders is their use for determining the position of satellites and spacecraft.

In last/latter decades were developed the diverse new radio-navigation systems: pulse, phase, frequency, etc. Despite the fact that some of these systems provide position finding with larger accuracy than the radio direction finders, last/latter completely retain their value. Is explained this to the facts that the radio direction finders possess a series of essential advantages. The radio direction finder, establish/installed aboard the aircraft or ship, can be used in any area, since do not require any special stations: direction finding can be produced on the constantly operating broadcast or communications radio stations. The equipment, adjustable aboard, is simple in operation and it is reliable. During the use of ground-based radio direction finders, not at all it is required any special equipment aboard of moving object, besides the normal receiving-transmitting station. The accuracy, provided with radio direction finders, is sufficient for the solution of the majority of navigational problems. Radio direction finder is the all-purpose instrument: it it is possible to utilize on large, medium and small distances, for example for the driving of aircraft on the predetermined course as booster agent for recovery to landing/fitting, etc. The advantage of radio direction finders is also simpler, than in the majority of other radio-navigation systems, the exchange of operating frequencies, which raises interference shielding, it makes it possible at each given torque/moment to select the frequency, least subjected to interferences. The work of radio direction finders is connected with short-term emission/radiation, and therefore does not cause the excessive charging of ether/ester.

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Finally, another navigation aids cannot replace radio direction finders in fields mentioned above of their application/use besides navigation (in military science, in scientific investigations, during the determination of direction aboard the ship, signalling of calamity).

In comparison with the radar methods of determining the position, the radio direction finders possess considerably larger range. It should be noted that goniometrical equipment/devices of radars are based actually on the same principles, as radio direction finders. However, as a result of the coordination of action of goniometrical equipment/devices with the remaining cell/elements of radar station, these equipment/devices possess a series of specific special feature/peculiarities. Therefore goniometrical equipment/devices of radar stations in this book are not examined.

Initially radio direction finders were fulfilled in the range of medium-frequency waves from that which is rotated by operator by antenna (framework) and with the auditory reading of bearing on the minimum of audibility.

Subsequently of development, they were directed toward operator's release from the rotation of antenna and the replacement of the auditory reading of bearing by reading along electromechanical instrument, along cathode-ray tube, in digital signal panel. Was carried out research on the reasons for the errors of radio direction finders, were found the measures of their elimination. Were developed the more effective antenna systems, ensuring large with accuracy and the sensitivity of radio traffic. Simultaneously was improved radio reception technique - were improved the indices of the forming part of radio direction finder receptors (interference shielding, sensitivity, the accuracy of installation and maintenance of frequency, reliability, etc.).

In connection with the common/general/total development of radio engineering and the mastery/adoption of high and ultrahigh frequencies, was expanded the frequency band of the work of radio direction finders. Contemporary direction-finding installations work in the range from the lowest to the highest frequencies.

As a result of the made investigations and developments, the direction finders are at present sufficiently precision instruments.

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However, the accuracy of direction finder is caused by a whole series of the factors, connected with it by itself and with the affect of the surrounding object/subjects, and also with the conditions of the propagation of electromagnetic waves on the way from transmitter to direction finder. Therefore the indicated high degree of accuracy can be provided only in that case when during the practical use of a radio direction finder is given up clear report in the processes, which occur in it, and in the different effects, exerted to these processes. Thus, for the proper practical use of a radio direction finder is necessary serious acquaintance with the theory of its work.

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Chapter 2.

PRINCIPLES AND METHODS OF RADIO TRAFFIC.

2.1. Electromagnetic field and its polarization.

It is known that electromagnetic field is the totality of the mutually connected electrical and magnetic fields.

Electromagnetic field of the radio waves, emitted by the transmitting antenna, is the field of the traveling wave: the phase of field varies in proportion to to the path of the propagation of wave, and amplitude changes relatively weakly.

Any emitter creates the induction fields and emission/radiation. At close distance from emitter (smaller than the wavelength) there are mainly fields of electrostatic and electromagnetic induction. The intensity/strength of the first inversely proportional to the cube of distance from emitter; the intensity/strength of the second inversely proportional to the square of distance. Radiation field is here relatively weak.

With distance from the emitter of induction field, rapidly they decrease at a distance, greater than two-three wavelengths, virtually remains one radiation field alone whose intensity/strength in free space inversely proportional to the first degree of distance of emitter.

In the zone of induction, in immediate proximity of emitter, between the strength of electrical and magnetic fields is a phase difference, close to 90°. With distance from transmitter, this phase difference decreases, and in the zone of emission/radiation during propagation in dielectric medium (air) electrical and magnetic fields coincide in phase.

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During the study of processes in radio direction finders us interests as radiation field when is oriented the distant radio transmitter, so also the field of the near zone when, for example, is investigated effect on the direction finding of the adjacent (to by antenna to the system of radio direction finder) metallic object/subjects.

At large distance from emitter (in the zone of emission/radiation) the vectors of the strength of electrical and magnetic fields (E and H) mutually perpendicular and perpendicular to Poynting's vector (S), which characterizes the direction of propagation of electromagnetic energy (Fig. 2.1). During free-space propagation constant-phase surfaces are the concentric spheres in center of which is located the emitter. This wave he is called spherical.

At large distances from the transmitting antenna the section of constant-phase surface near observation point can be considered plane, i.e., to consider wave as plane.

The structure of field is distorted near the interfaces (for example, the earth/ground and air), and also in the presence of any obstructions or secondary emitters: mountains, trees, antennas, etc.

The vectors of the strength of electrical and magnetic fields, remaining mutually perpendicular, can have different direction. For the characteristic of the sense of the vector of field, is introduced the concept of polarization. The polarization of electromagnetic field he is called the orientation of the electric field of wave relative to the plane of propagation. The plane of propagation he is called the plane, which contains direction of propagation and perpendicular to the earth's surface.

Can be observed the following forms of polarization:

1. Normal, or is vertical, the polarization, when the vector of electric field lie/rests at the plane of propagation. This case is depicted on Fig. 2.2, where zy is a vertical plane.

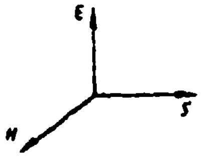

Fig. 2.1. Mutual location of vectors of electromagnetic field.

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2. Abnormal polarization when vector of electric field composes certain angle with vertical earth referenced plane, which contains direction of propagation. A special case of abnormal polarization is horizontal polarization, when the vector of electric field is horizontal, and the vector of magnetic field is vertical.

The indicated in p. 1 and 2 polarizations they are linear. During any linear polarization electric field can be decomposed on two fields - vertical and horizontal, cophasal.

3. If between vertical and horizontal components of electric

field is phase displacement, then is obtained resulting eliptically polarized field. The terminuses of the vector of directivity of field during the period of high frequency describe ellipse. The rotation of the vector of field in time is realized unevenly. Direction of rotation depends on a phase difference the vertical and horizontal components of electric field. A special case of elliptical polarization is circular polarization, when vertical and horizontal components are equal to each other and a phase difference is equal to 90°.

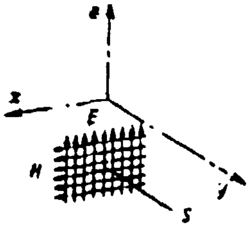

Fig. 2.2. Normal polarized electromagnetic field.

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The strength of the electrical and magnetic fields of plane electromagnetic wave in the general case of elliptical polarization it is possible to express by the formulas:

$$\left.\begin{aligned}\vec{E}&=(\vec{u}_0 E_u+j\vec{v}_0 E_v)\,e^{j\Phi}e^{-jmr\cos\beta\cos(\theta-\theta)},\\ \vec{H}&=(-j\vec{u}_0 H_u+\vec{v}_0 H_v)\,e^{j\Phi}e^{-jmr\cos\beta\cos(\theta-\theta)},\end{aligned}\right\}\qquad(2.1)$$

where $\vec{u}_0$ and $\vec{v}_0$ - the unit vectors, which characterize the direction of the large and minor axes of the ellipse of the polarization of electric intensity;

E_u, E_v, H_u, H_v — value of the large and semiminor axes of the ellipse of the polarization of electrical and magnetic fields (Fig. 2.3);

m - wave number ($m = 2\pi/\lambda$ in free space);

r and φ - the polar coordinates of the point in question, relative to the origin of coordinates;

ψ — the phase in the beginning of coordinates;

θ - the angle of the direction of propagation of wave with initial reference line (azimuth, bearing);

β - the angle of the slope of a front of wave (Fig. 2.4).

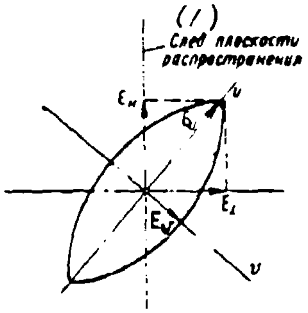

Fig. 2.3

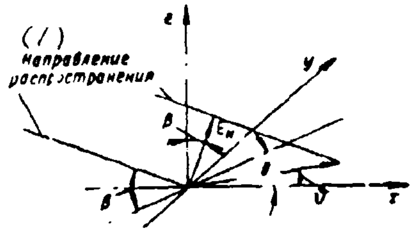

Fig. 2.4

Fig. 2.3. Elliptically polarized electric field.

Key: (1). Trace of propagation.

Fig. 2.4. Adopted designations of coordinates.

Key: (1). Direction of propagation.

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Elliptically polarized wave can be presented as sum of two linearly polarized waves: in the plane of propagation $(E_n, H_\perp)$ it is perpendicular to it $(E_\perp, H_n)$

$$\vec{E} = (\vec{n}E_n + \vec{a}E_\perp e^{j\psi_a})\, e^{j\psi} e^{-jmr\cos\beta\cos(\theta-\theta)}, \qquad (2.2)$$

$$\vec{H} = (-\vec{n}H_\perp e^{j\psi_a} + \vec{a}H_n)\, e^{j\psi} e^{-jmr\cos\beta\cos(\theta-\theta)}, \qquad (2.3)$$

where $\vec{n}$ is the unit vector, which lies at the plane of propagation and perpendicular to direction of propagation;

$\vec{a}$ - the unit vector, perpendicular to the plane of propagation;

E_n, $E_\perp$, H_n, $H_\perp$ — the corresponding components of electrical and magnetic field;

ψ_a — phase displacement between the normally and abnormally polarized components.

The angle, composed by the direction of the transverse of the polarization of electric field with the plane of propagation, he is called the angle of polarization γ.

The component of electric field, which lies at the plane of propagation E_n can be decomposed on two components: vertical E_z and horizontal E_x in the plane of propagation. If θ = 0, i.e., if direction of propagation coincides with X-axis, we will obtain following components along the axes of the coordinates:

vertical

$$E_z = \cos\beta \sqrt{(E_u \cos\gamma)^2 + (E_v \sin\gamma)^2} = E_n \cos\beta;$$

horizontal in direction of propagation

$$E_x = \sin\beta \sqrt{(E_u \cos\gamma)^2 + (E_v \sin\gamma)^2} = E_n \sin\beta;$$

horizontal, perpendicular to direction of propagation

$$E_y = \sqrt{(E_u \sin\gamma)^2 + (E_v \cos\gamma)^2} = E_\perp.$$

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2.2. Principles of radio traffic.

Examining expressions for the strength of the field of plane wave (2.2) and (2.3), we see that the sense of the vector of field and the phase of the strength of field depend on the angle of arrival of wave. the use of these dependences makes it possible to carry out radio traffic.

For determining the sense of the vector of the fields, it is possible to utilize an electrical or magnetic dipole, practical fulfillment of which is the short vibrator or the framework of small size/dimensions. Revolving dipole, we will obtain the maximum of emf in it, when its axis coincides with the direction of the transverse of the polarization of the strength of electrical or magnetic field for electrical and magnetic dipoles respectively. The minimum of emf will be obtained, when the axis of dipole is parallel to the minor axis of the ellipse of polarization. However, the direction of the axes of the ellipse of polarization depends not only on direction of propagation, but also on the angles of the slope of a front of wave and on the angle of polarization.

Only if the angle of the slope of a front of wave θ is equal to zero either the angle of polarization γ it is equal to 0° or 90°, is possible error-free direction finding with the aid of dipole. In the first case ($\beta = 0$) the direction of propagation coincides with perpendicular to the horizontal axis of the ellipse of polarization. In the second case ($\gamma = 0$ or $\gamma = 90°$) the direction of propagation coincides with perpendicular respectively to the vector of the magnetic or electric field of electromagnetic wave.

Research on radiowave propagation is led to the conclusion that under specific conditions (for example, during the propagation of the terrestrial wave above the well conducting surface) is provided by an angle of polarization, close to zero. In this case, the vector of electric intensity is vertical, the vector of magnetic intensity horizontal and for direction finding can be used magnetic dipole, i.e., the framework. Perpendicular to the axis of dipole with the minimum of signal coincides with direction of propagation. An antenna of this type finds wide application, especially on the medium and long waves with which the polarization in many instances is normal. It is necessary to note that during appearance, on the strength of the conditions of propagation, abnormal polarization the direction finding to the framework will be accompanied by the errors which are called polarizational and in more detail they are examined in chapter 6.

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Stable polarization at an angle of 90° ($\gamma = 90°$), when electric intensity is horizontal, is observed relatively rarely. Only in these cases electrical horizontal dipole can be used for radio direction finding.

The phase of the strength of field also depends on parameters enumerated earlier (azimuth, the angle of the slope of a front of wave, angle of polarizations) whose separate determination according to observed data in a single point is impossible. The measurement of the phase at several points, which requires the application/use of the diversity-reception antennas, makes it possible to compose the system of equations whose solution gives the values of all indicated parameters of electromagnetic field.

The number of parameters and at the same time the number of necessary measurements decreases, if we determine the strength of the field of the determined polarization. For this purpose, sensing devices of radio direction finders are designed for the reception/procedure only of one of field components, usually vertical components of electric field.

The exception/elimination of the reception/procedure of the second (horizontal) component of electric field technically complicatedly and virtually entirely is not reached. The remanent/residual reception/procedure of the horizontal component of electric field leads to the errors which are also called polarizational. One should emphasize the difference in the reasons for polarizational errors with direction finding with the aid of the framework and with the aid of the spaced antennas: in the first case the polarizational errors are characteristic to operating principle, in the second they are the consequence only of the inadequacy of fulfillment.

In this chapter we will consider the action only of vertical component of the electric field

$$\dot{E} = E_n e^{j\psi} e^{-jmr\cos\beta\cos(\theta-\Phi)}. \qquad (2.4)$$

In many instances the propagation of wave from transmitter has the multiple-pronged character: besides direct wave are propagated the waves, reflected from the different layers or heterogeneity of ionosphere or the troposphere, and also the wave, reflected from the earth/ground, from different return emitters, which are located on greater to silt smaller distance from direction finder and, etc.

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The resulting interference field is sum N of the coherent waves

$$\dot{E}=\sum_{i=1}^{N} E_{ni}e^{j\psi_i}e^{-jmr\cos\beta_i\cos(\theta_i-\theta)}. \qquad (2.5)$$

Each wave is characterized four by independent parameters $(E_{ni}, \psi_i, \beta_i, \theta_i)$, and interference field as a whole by 4 N parameters. Amplitude and the phase of field of one of the waves can be accepted arbitrary for the comparison with them of other waves. then it is required to determine 4N-2 the unknowns of the parameter. Each antenna provides information about amplitude and the phase by induced in it emf. are possible only relative measurements of amplitude and phase relative to amplitude and phases of one of the antennas. Therefore, having n of antennas, we obtain 2(n - 1) the results of measurement. Knowing the locations of antennas, we can be 2(n - 1) the equations, which relate the parameters of waves with amplitudes and phases of the stresses in antennas. Equalizing the number of unknowns to the number of equations, we obtain

$$2(n-1)=4N-2 \qquad (2.6)$$

or

$$n=2N.$$

For the separate determination of the parameters of all incident waves, the number of antennas must be equal to the doubled number of waves. This direction finder will ensure the error-free determination of direction and angle of incidence in each of the component waves. For the production of radio direction finder according to this principle besides by the antenna of system from the necessary number of antennas are required the very complex receiving and computers, intended for the solution to the indicated equations. For this reason up to now, there is no radio direction finder, which in practice realizes the given operating principle. All the existing radio direction finders according to operating principle are designed for the direction finding of one wave. When on this "single-wave" radio direction finder operates the composite field of several waves, appear the errors, called interference and examine/considered in chapter 6.

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In single-wave radio direction finder the required number of antennas is greater than determined by formula (2.6). This is explained to the facts that with the incidence of only wave the amplitude of the strength of field and, therefore, amplitude of emf in antennas in all measuring points are identical and the data on them cannot be used. We obtain possibility to comprise only $(n - 1)$ equations according to the results of the relative measurements of phase. The number of unknowns is equal to two (θ and β). Hence it follows that for the direction finding of one wave ($N = 1$) is required not less than three motionless antennas. The number of antennas of radio direction finder frequently exceeds the theoretical minimum. Surplus antennas make it possible to simplify the technical fulfillment of radio direction finder and to improve its indices: sensitivity, interference shielding, instrument/tool accuracy, etc.

Since in the process of direction finding is usually of interest only azimuth of incident wave, it is possible to search for the method to decrease the required number of antennas, excluding the possibility of the determination of high-altitude angle β. From formula (2.4) it is evident that, accepting the phase of field in one of the antennas of the equal to zero and revolving the second antenna, it is possible to find this value of the angle $\vartheta = (\varphi = \pi/2 + \theta)$ at which and the phase of field for the second antenna is turned into zero independently angle of the slope of a front of wave β. In this case of the sufficiently two rotatable antennas for the determination of bearing; however, without the possibility of determining the high-altitude angle. Direction finding actually is reduced to the determination of the direction of the lines of the identical phases of the field perpendicularly to which is arrange/located transmitter.

The angle of the slope of a front of wave under normal conditions is changed within not wide limits near 0. If we consider angle β known, after accepting for it certain average value, is possible direction finding to two motionless antennas, since the number of unknown parameters is reduced to one θ. However, the actual value of angle β can differ from that which was accepted during calculation. Because of this appear the errors in the determination of bearing, which obtained the designation of high-altitude.

We examined the field of single electromagnetic wave with the stable parameters.

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Under the actual conditions of radiowave propagation with reflection from "rough" ionospheric layers or due to the effect of the troposphere the value and the phase of field, its polarization, the angle of the slope of a front of wave and the direction of arrival vary around average value. A wave of such type can be considered as wave, which has the angular spectrum, i.e., as totality or the beam of waves with the different directions, which are usually placed within the limits of small angle. Since radio direction finders of the type examined above cannot solve separate component waves, their readings correspond to the result of the effect of all elementary waves, i.e., give errors. Errors oscillate in the course of time in accordance with the fluctuations of parameters of complex wave.

Analogous errors are characteristic to the multiwave radio direction finder which solves separate discrete waves, but it cannot solve the continuous angular spectrum of scattered beam. Errors of this type also are related to interference.

After feed/conducting result, it is possible to establish that according to operating principle the radio direction finders are subdivided into the following groups:

1. Radio direction finders with the single rotatable dipole (virtually usually framework). These radio direction finders are subjected to polarizational errors.

2. Radio direction finders with two rotatable antennas.

3. Radio direction finders with two motionless antennas. These radio direction finders are subjected to high-altitude errors.

4. Radio direction finders with single-wave type three or more motionless antennas.

All the enumerated direction finders are subjected to interference errors.

5. Multiwave type radio direction finders.

2.3. Direction-finding methods.

During use for determining the bearing of several antennae arrange/located at the diverse points, the information about bearing is contained in the phases of the strength of field and, therefore, in the phases of emf, induced in antennas. It is possible to distinguish two methods of processing this information.

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With the first method of the voltage of separate antennas of up to supply to the input of receiver-amplifier device, they are combined so that either the carrier amplitude of the resulting stress or the parameters of the amplitude modulation (depth or the phase of amplitude modulation) of the resulting stress are the function of bearing. The determination of bearing is conducted by the measurement of amplitude or parameters of the amplitude modulation of the output voltage of receiving indicator. The radio direction finders, which operate using this method, they are called amplitude. Another method of determining the bearing is based on the application/use of phase measurements. The radio direction finders in which is used the measurement of the phase of high-frequency oscillations, they are called phase.

Amplitude direction-finding methods. Let us examine the application/use of amplitude methods in the case the use of the rotatable antenna with (framework, two diverse framework, two and more spaced antennas, the combinations of the framework, and antennas). The resulting voltage/stress by the antenna of system depends on the direction of incident wave, i.e.,

$$U = \mathrm{E} f(\theta, \beta). \qquad (2.7)$$

The dependence of output potential by the antenna of system from the direction of the arrival of wave he is called directional characteristic of antenna, and its graphic representation - by radiation pattern. Is is commonly used standardized/normalized directional characteristic

$$F(\theta, \beta) = \frac{f(\theta, \beta)}{f_{\text{макс}}(\theta, \beta)}, \qquad (2.8)$$

where $f_{\text{макс}}(\theta, \beta)$ — the maximum value of directional characteristic.

Figures 2.5 and 2.6 as an example gives different radiation patterns in horizontal plane (i.e. with $\beta = 0$), presented in Cartesian and polar coordinates. Figures 2.5a depicts in polar coordinates widespread directional characteristic

$$F(\theta) = \cos\theta. \qquad (2.9)$$

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In form of image this characteristic calls the "diagram of eight". On Fig. 2.5b, this same the characteristic is depicted in rectilinear coordinates.

On Fig. 2.6a and b, is depicted acute/sharper radiation pattern in polar and rectilinear coordinates.

For determining bearing antenna system they revolve, observing output voltage. It is possible to distinguish two methods of the rotation:

1) the rotation of antenna of up to obtaining of minimum or maximum of output voltage, after which it is conducted the reading of bearing on antenna position;

2) the long running of antenna.

Let us call/name the first method mounting method on bearing.

Let us disassemble it in more detail. The rotation of antenna can be realized by hand by the operator who finds bearing on audibility in telephone (auditory direction finding) or from readings of visual display (visual nonautomatic direction finding).

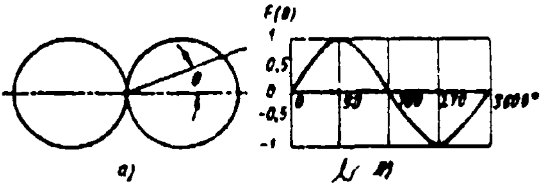

Fig. 2.5. Cosinusoidal directional characteristic.

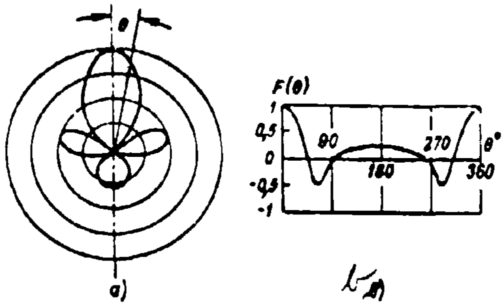

Fig. 2.6. Directional characteristic.

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As indicator are applied the dial instrument or the cathode-ray tube. Rotation can be realized also automatically with the aid of the power drive, controlled by the output voltage of radio direction finder.

In work for audition determination of direction, they produce on the minimum (disappearance) of audibility in telephone, but not on maximum, since the first method is considerably more precise. Actually, with direction finding with respect to the minimum, the very small (order 0.5-3°) divergence of antenna from the position of zero reception/procedure already causes the appearance of completely noticeable audibility in telephone. Near maximum the audibility changes much more slowly and is possible sufficiently considerable divergence from the position of the true maximum of reception/procedure before we will note a change in sound intensity. Let us examine, for example, the use of cosinusoidal directional characteristic. It is experimentally establish/installed that on the middle ear notes a change in sound intensity. Let us examine, for example, the use of cosinusoidal characteristics of directivity. It is experimentally establish/installed that on the middle ear notes a change in the audibility to 7-8o/o. This difference we will obtain near the maximum when cos θ becomes equal to 0.92-0.93, which corresponds to angle of 23-21°. Consequently, having only gone away from maximum to the angle, greater than 20°, we will note a change of the audibility in telephone.

Besides the determination of bearing for audition from the minimum of radiation pattern, there is another auditory method, based on the comparison of audibility in two positions by the rotatory antenna of system or during switching by the antenna of system, combined with the omnidirectional antenna.

The simplest diagram of the realization of comparison method is represented on Fig. 2.7. Two Framework - fundamental framework A and auxiliary B are arrange/located mutually perpendicularly and are attached on one axis. Emf from framework B of switch by switch D. Direction finding consists in the rotation of both framework and in the determination of such position by which switchings of switch D do not change intensity of reception.

On Fig. 2.8 dotted lines, gave the radiation patterns of framework A and B. As heavy and fine/thin solid lines are depicted total radiation patterns in two positions of switch D.

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According to the same law will change the audibility at the output of receiver during the rotation of the framework. The position of the framework, in which the indicated switchings do not change audibility, corresponds to direction 3-4, when the plane of framework B to the perpendicular of the location of radio transmitter (1-2). Then there is no reception/procedure to framework B. The direction of the incoming electromagnetic wave is determined in this case by perpendicular to the plane of framework B. On the dial/limb of radio direction finder, must be counted off the angle between the initial reference line and this perpendicular. Audibility will not change also in the position of the framework along directions 1 and 2, i.e., in the absence of reception/procedure in framework A. this could lead to the possibility of bearing error to 90°. Virtually this ambiguity is removed by the facts that framework B are taken with the larger effective height than A; therefore audibility with accurate bearing is less than with erroneous bearing. Framework B one should take greater size/dimensions than framework A, and for an increase in the accuracy of reading (see § 2.7).

Instead of the auxiliary framework B, it is possible to use the omnidirectional antenna whose emf coincides in phase with emf from the framework. Emf of the framework or antenna is changed over during direction finding. Direction finding consists in the determination of the position of the framework, by which the audibility at the output of receiver is not changed during switchings.

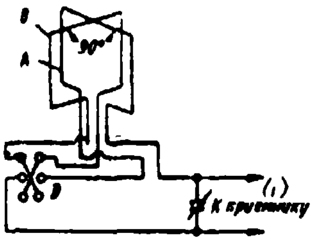

Fig. 2.7.

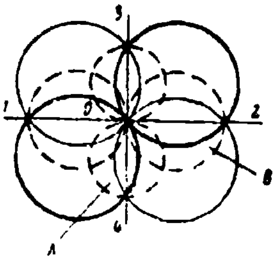

Fig. 2.8.

Fig. 2.7. Diagram of the realization of comparison method.

Key: (1). To receiver.

Fig. 2.8. Radiation pattern during switching of the framework.

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Between the amplitudes of emf of the framework and antenna, there can be in principle any relationship/ratio. Let us assume that the value of emf of the changed over antenna is equal to maximum emf of the framework. Then the resulting radiation pattern will be obtained as on Fig. 2.9, on which by fine/thin solid line is shown the antenna radiation pattern, by fine/thin dotted line - the radiation pattern of the framework. By heavy broken and solid lines are shown the resulting diagrams in two positions of switch D. As can be seen from Fig. 2.9, the value of resulting emf is not changed during switchings D, when the framework is directed lengthwise Oa and Ob, i.e., when the plane of the framework is perpendicular to direction in radio station. This is considered during the orientation of the framework and dial/limb in the indicated system.

During manual unit on bearing according to instrument (visual for nonautomatic unit is most expedient to utilize comparison method. The voltage, obtained as a result of switchings, is modulated. Utilizing, for example, radiation pattern of the combination of the antenna and framework, as on Fig. 2.9, we see that in position Oa the voltage during one half-period of switching will be proportional 04, and during the second - is proportional 03. This is represented on Fig. 2.10. Radio frequency voltage is modulated and has rectangular envelope. In position Ob (Fig. 2.9) the depth of modulation is equal to zero, since voltages 01 into both half-periods switchings are equal to each other. During the divergence of antenna to the position, symmetrical Oa, we will obtain the same depth of modulation as in Oa, but with the inverted phase. It is easy to see that the depth of modulation is the function of angle of rotation. Unit with the antenna of system on bearing produces on minimum (zero) of the depth of modulation.

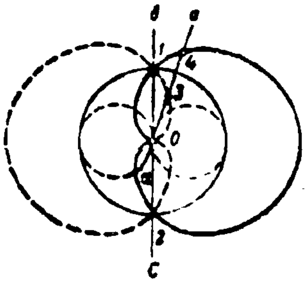

Fig. 2.9. Radiation pattern of system from the framework and the open antenna.

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In visual radio direction finders at the output of receptor, is included the indicator according to readings of which the operator determines the antenna position, which corresponds to the minimum of the depth of modulation. In the automatic direction finders the follower with the aid of the power drive turns antenna in the position, which corresponds to zero modulation. Besides square-wave modulation can be obtained sinusoidal modulation.

Direction finding on maximum, as was noted above, with cosinusoidal radiation pattern is imprecise. However, with sufficiently acute/sharp radiation pattern, This method of direction finding becomes possible. Specifically, on ultra short waves it is easy to perform antenna system with this acute/sharp radiation pattern that the direction finding on maximum will be sufficiently precise.

Let us examine the second method - with the long running of antenna. In the radio direction finders, working using this method, the voltage proves to be modulated depending on the frequency of the rotation of antenna. Actually, if angular frequency Ω, then

$$\vartheta = \Omega t$$

and the voltage of antenna is proportional to directional characteristic

$$F(\vartheta - \theta) = F(\Omega t - \theta),$$

i. e.

$$u = U_{\text{макс}} F(\Omega t - \theta). \qquad (2.10)$$

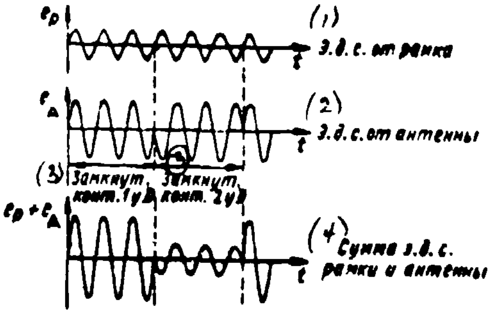

Fig. 2.10. Addition of the voltages of the framework and antenna.

Key: (1). Emf from the framework. (2). Emf from antenna. (3). They will close cont. (4). Sum of emf of the framework and antenna.

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Expression (2.10) represents modulated voltage, moreover the phase of modulation curve is determined by bearing θ. At the output of receptor, is switched on the instrument according to which is counted off the phase of modulation frequency, which corresponds to bearing. Thus, radio direction finders with the long running of antenna are automatic, i.e., making it possible to directly count off bearing without observer's any operations. They are called phase-meter.

Let us turn now to radio direction finders with motionless antennas. Let us examine the simplest antenna system of four vertical wire antennas [1] (Fig. 2.11).

Antennas 1, 2, 3, 4 are arrange/located in the apex/verteres of square, direction of one of the diagonals of square with antennas 1-3 coinciding with the initial reference line of bearing.

Let us designate: h_c - acting height of antenna; 2b - the separation of opposite antennas (corner diameter); ψ the initial phase of field at the center of system.

Emf, induced in antennas, in accordance with (2.4) will be

$$\left.\begin{aligned} \dot{U}_1 &= Eh_c e^{j(\psi + mb\cos\theta\cos\beta)}, \\ \dot{U}_2 &= Eh_c e^{j(\psi - mb\sin\theta\cos\beta)}, \\ \dot{U}_3 &= Eh_c e^{j(\psi - mb\cos\theta\cos\beta)}, \\ \dot{U}_4 &= Eh_d e^{j(\psi + mb\sin\theta\cos\beta)}. \end{aligned}\right\} \qquad (2.11)$$

FOOTNOTE [1]. The application/use of four antennas instead of three antennas, the minimum number, simplifies examination. As has already been indicated that for an improvement in the series of the indices of direction finder, is applied a larger number of antennas. ENDFOOTNOTE.

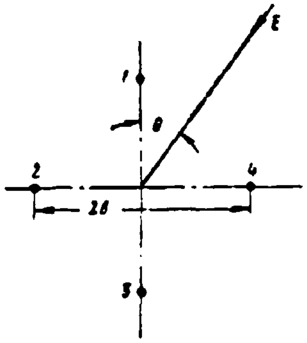

Fig. 2.11. Antenna location.

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The differential voltages of opposite antennas to sex/floor are expected equal to

$$\dot{U}_{13} = \dot{U}_1 - \dot{U}_3 = j2Eh_e e^{j\phi} \sin(mb\cos\beta\cos\theta), \quad (2.12)$$
$$\dot{U}_{24} = \dot{U}_2 - \dot{U}_4 = j2Eh_e e^{j\phi} \sin(mb\cos\beta\sin\theta). \quad (2.13)$$

We obtain two voltages of the identical phase whose amplitude depends on bearing θ and on the angle of the slope of a front of wave β. With the aid of the appropriate computer it is possible to determine by the obtained two voltages both angles θ and β. Virtually in this type, radio direction finders is applied a small separation of antennas in comparison with wavelength ($2b < \lambda$), since with a small separation is simplified the procedure of the determination of bearing. Furthermore, with large separation the solution to equations (2.12) and (2.13) proves to be many-valued. Taking into account a small separation, in both expressions sines can be replaced with arguments. Then

$$\dot{U}_{13} = j2h_{e}Ee^{j\psi}mb\cos\beta\cos\theta, \qquad (2.12')$$
$$\dot{U}_{24} = j2h_{e}Ee^{j\psi}mb\cos\beta\sin\theta. \qquad (2.13')$$

Replacement of sines by arguments, strictly speaking, it is permissible only with the negligibly low values of arguments. During an increase in the arguments of expression (2.12'), (2.13') they become imprecise and the obtained on them subsequently bearing is accompanied by the error which is called the error of separation. The maximum permissible errors of separation limit the distance between antennas in radio direction finders of the type in question.

For determining angle θ in accordance with (2.12'), (2.13') can be used two methods. The first method consists in the fact that between antenna system and the receiver is included the goniometer, which consists of two motionless field coils and one revolving search coil.

Antennas 1-3 and 2-4 are connected with the aid of feeders to two motionless mutually perpendicular field coils I, of II goniometer. Within field coils rotates third search coil of goniometer, connected to the input of receiver.

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The appearing under the effect of the differential voltages U_{13} and U_{24} currents and created by them in field coils I and II magnetic fields H_I and H_{II} are proportional to differential voltages, i.e.,

$$H_I = k_1\dot{U}_{13} = H_{I\,\text{макс}}\cos\theta,$$
$$H_{II} = k_2\dot{U}_{24} = H_{II\,\text{макс}}\sin\theta,$$

where k_1 and k_2 the proportionality factors, which depend on the parameters of goniometer and antennas.

The resulting magnetic field in goniometer is equal to vector sum of fields H_I and H_{II}. Let us assume that the currents in both coils I and II, and also the created by them magnetic fields H_I and H_{II} are located in phase. Then the vector of the resulting magnetic field in value and direction will be determined by the diagonal of rectangle (Fig. 2.12).

Magnitude of vector the resulting magnetic field is expressed

$$H = OR = \sqrt{OA^2 + OB^2} = \\ = \sqrt{H^2_{I\,\text{макс}}\cos^2\theta + H_{II\,\text{макс}}\sin^2\theta}, \qquad (2.14)$$

and its direction is determined by the angle Φ of vector H with standard to the plane of first field coil I, which coincides with magnetic field H_I, moreover

$$\mathrm{tg}\,\Phi=\frac{OB}{OA}=\frac{H_{II\,\text{макс}}\sin\theta}{H_{I\,\text{макс}}\cos\theta}. \qquad (2.15)$$

Let us make the further assumption that

$$H_{I\,\text{макс}}=H_{II\,\text{макс}}=H_{\text{макс}}.$$

Then according to (2.14) and (2.15) we obtain that the resulting magnetic field $H=H_{\text{макс}}$ does not depend on angle θ and that

$$\mathrm{tg}\,\Phi=\mathrm{tg}\,\theta \quad \text{or} \quad \Phi=\theta.$$

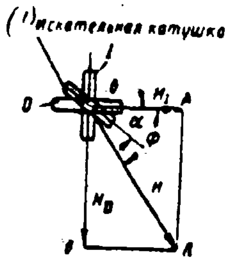

Fig. 2.12. Addition of magnetic fields in goniometer.

Key: (1). Searching coil.

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Consequently, when making these assumptions the magnetic field strength within goniometer on depends on the direction of incident wave. Direction of the magnetic field composes with standard to the plane of the first field coil of I goniometer accurately the same angle Φ, which composes the direction of the arrival of wave with the plane of antennas 1-3.

If one assumes that field within goniometer uniform, then of emf $E_{\text{и}}$, induced in search coil, will be proportional to the resulting field H, multiplied by sin ($\Phi - \alpha$), where α is the angle between the standard to the plane of the first field coil and the plane of search coil, calculated according to the scale of goniometer; $(F - \alpha)$ - the angle between the plane of search coil and the direction of the resulting magnetic flux (Fig. 2.12), i.e.

$$E_{\text{п}} = kH\sin(\Phi - \alpha), \qquad (2.16)$$

where k - the proportionality factor, depending on the parameters of goniometer.

In goniometric to system instead of the pairs of spaced antennas 1-3 and 2-4 it is possible to use the framework. Without being stopped here on the theory of the action of goniometer which is examined in detail in chapter 3, let us note that zero output voltage are obtained from (2.16) during the rotation of search coil through angle $\alpha = \theta$. This position of search coil gives directly the reading of bearing. The process of direction finding in goniometric system is obtained by the same as with direction finding by mounting method on the bearing of the rotatory antenna. It is possible to also carry out long running of goniometer, applying for the indication of bearing the same reception/procedures as with that which is rotating antenna. In other words, goniometer makes it possible to carry out rotation of cosinusoidal radiation pattern with motionless antennas.

With the other method of determining the bearing on the basis of the equations (2.12'), (2.13') of voltage U_{13} and U_{24}, they are amplified in two independent receivers (channels), the factor of amplification of which is differing (by value and phase). The intensive voltages will be feed/conducted to plate X and Y of cathode-ray tube (Fig. 2.13). The trace of electron beam will describe the straight line whose slope/inclination is equal to

$$\operatorname{tg} \alpha = \frac{Y}{X} = \frac{U_{24}}{U_{13}} = \operatorname{tg} \theta, \quad e \quad \alpha = \theta.$$

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The position of the glowing line on the screen of cathode-ray tube determines directly bearing. This type radio direction finders they are called two-channel. They are automatic. It should be noted that it is possible to carry out independent amplification of two voltages in one receiver, applying the frequency, phase or time sharing of the voltages of channels.

In the radio direction finder with two motionless antennas, constructed according to amplitude principle, is utilized usually the sum-and-difference method of reading [1].

---

[1] FOOTNOTE [1]. The sum-and-difference method of reading sometimes is related to phase-difference direction-finding method. ENDFOOTNOTE.

Differential voltage from one pair of antennas, arrange/located, for example, as antennas 1 and 3 on Fig. 2.11, it will be

$$\dot{U}_{\Delta}=\dot{U}_{13}=j2Eh_{e}e^{j\varphi}\sin(mb\cos\beta\cos\theta). \qquad (2.17)$$

Let us find also the total voltage

$$\dot{U}_{\Sigma}=\dot{U}_{1}+\dot{U}_{3}=2Eh_{e}e^{j\varphi}\cos(mb\cos\beta\cos\theta). \qquad (2.17')$$

Each of these voltages will be feed/conducted to independent receiver-amplifier channel. Both channels have identical amplification factors. In one of the channels, the voltage undergoes phase displacement 90°.

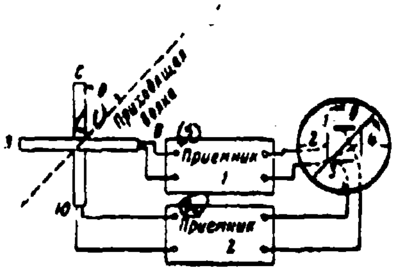

Fig. 2.13. Diagram of two-channel radio direction finder.

Key: (1). Incident wave. (2). Receiver.

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At the output of channels, we will obtain the voltages of the identical phase, proportional U_Σ and U_Δ. Voltages will be feed/conducted to plates $X(U_\Sigma)$ and $Y(U_\Delta)$ cathode-ray tube. Inclination of the glowing line on tube face is equal to

$$\mathrm{tg}\,\alpha = \frac{Y}{X} = \mathrm{tg}\,(mb\cos\beta\cos\theta)$$

or

$$\alpha = mb\cos\beta\cos\theta.$$

If we count off angles not from the line of antenna location, but from standard to it, then $\alpha = mb\cos\beta\sin\theta'$.

Angle α is connected with bearing θ'. If one assumes that the angle β is known, then on angle α it is possible to find θ'.

The application/use of a small separation of antennas, component motionless pair, is inexpedient, since the rotating antenna couple or goniometric system are simpler in operation. Use by a motionless antenna vapors acquires sense with large separation, since in this case is raised the accuracy of direction finding.

For the possibility of the realization of single-valued reading, the angle α must not be more $\pi/2$, i.e.,

$$mb \sin \theta' < \frac{\pi}{2}. \qquad (2.18)$$

Last/latter condition with that which was assigned b/λ limits the sector within limits of which it is conducted direction finding. From expression (2.18) at the low values of angle θ' the sector of direction finding is determined from formula $2\theta' < \pi/2mb$. Thus, this direction finder is sector unlike the direction finders of other types, which allow/assume direction finding within limits of 360°. Taking into account this property, it is expedient each of two antennas to fulfill with acute/sharp radiation pattern, as a result of which is raised the interference shielding of direction finder, are lowered interference errors it decreases the probability of gross error due to multiformity.

A. low values the angle θ' can be designed on the obtained angle α by simple indexing into conversion factor $k_n = mb\cos\beta$, larger than unity. It is easy to see that the conversion factor is led to an increase in the accuracy of direction finding.

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In conversion factor enters the angle of the slope of a front of wave. If in calculation is undertaken the value of angle β, different from its unknown to us actual value, in the determination of bearing we will obtain high-altitude error that it corresponds to the overall considerations, expressed in § 2.2.

Phase-difference direction-finding methods.

Let us examine the application/use of phase-difference methods in radio direction finders with the rotatable antennas. Let one of them (motionless) be arrange/located in the beginning of coordinates, and the second (rotating) - at a distance r from the first antenna at an angle ϑ to initial reference line.

The strength of field and the proportional to them voltages in two antennas in question will be

$$\dot{U}_1 = h_e \dot{E}_1 = h_e E e^{j\varphi},$$
$$\dot{U}_2 = h_e \dot{E}_2 = h_e E e^{j\varphi} e^{-jmr\cos\beta\cos(\vartheta-\theta)}.$$

A phase difference of stresses U_1 and U_2 is equal to

$$\psi_p = mr\cos\beta\cos(\vartheta - \theta).$$

In this type, direction finders is utilized the dependence of phase on the angle of bearing. For the determination of this dependence, the second antenna rotates on radius r with angular frequency Ω, so that

$$\vartheta = \Omega t.$$

The phase of stress U_2 relative to voltage U_1

$$\psi_p = mr\cos\beta\cos(\Omega t - \theta) \qquad (2.19)$$

proves to be that which is changing in time. Consequently, differential voltage is modulated on phase. It is known that phase modulation can be examined just as frequency. The deviation of frequency is equal to

$$\Delta\omega = \frac{d\psi_p}{dt} = -mr\Omega\cos\beta\sin(\Omega t - \theta). \qquad (2.20)$$

The phase both of phase and frequency modulation corresponds to bearing.

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Receptor (Fig. 2.14) must contain high-frequency amplifier, the phase either FM discriminator, which isolates the low-frequency voltage, proportional to changes in phase (2.19) or of frequency (2.20), low-frequency amplifier and the phase indicator from readings of which is counted off the phase of output voltage, which corresponds to bearing. This type direction finders they are called direction finders with the cyclic measurement of phase in high frequency. Instead of the rotatable antenna can be applied the motionless changed over antennas (see § 8.8).

In the case of motionless antennas, it is possible to utilize antenna location in accordance with Fig. 2.11.

The voltage of each of the antennas individually is amplified in independent receiver-amplifier channels, with accurately identical phase responses. Is conducted the measurement of a phase difference in output potentials first and second channels ψ_{13} and on the output of the third and fourth channels ψ_{24}. From expressions (2.11) it follows that

$$\psi_{13} = 2mb\cos\beta\cos\theta \qquad (2.21)$$

and

$$\psi_{24} = 2mb\cos\beta\sin\theta. \qquad (2.22)$$

According to these data can be found the angle θ (and also angle β). Let, for example, in diagram be developed the voltages, proportional to phases $U_x = \psi_{13}$ and $U_y = \psi_{24}$.

Let us conduct these voltages to plate X and Y of cathode-ray tube. The angular position of focus is determined by angle α, moreover

$$\operatorname{tg}\alpha = \frac{Y}{X} = \frac{U_y}{U_x} = \frac{\psi_{24}}{\psi_{13}} = \operatorname{tg}\theta.$$

Consequently, angle $\alpha = \theta$ directly determines bearing. Can be developed the diagram with the aid of which is determined also the angle β.

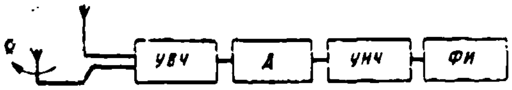

Fig. 2.14. Diagram of radio direction finder with the cyclic measurement of phase.

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If we are restricted to the application/use only of two motionless antennas (for example, 1 and 3), available it will remain only one of the equations (2.21) or (2.22), from which it is possible to determine θ, by assigning angle β. It is natural that the determination θ is accompanied by high-altitude errors, if assigned imprecise value of angle β.

With the large distance between antennas, appears the multiformity.

In radio direction finders of this type for the target/purpose of an increase in the accuracy, is applied usually the large separation of antennas, and multiformity they solve by the application/use of the second system (phase or amplitude) with a small separation, that gives the single-valued (but less precise) reading of bearing (§ 8.7).

According to this same principle it is possible to construct the "multiwave" radio direction finder, which allows separate direction finding of several (with four antennas of two) coherent waves, which operate on it simultaneously. As already mentioned that in this case strongly becomes complicated receiving-indicator equipment/device [2.15].

Above are examined two fundamental direction-finding methods: amplitude and phase. On the first of them, the phase dependences of electromotive forces in antennas on bearing are converted into amplitude (radiation pattern) at the input of receiver. According to the second method, phase, phase relationship/ratios are retained before leaving of receiver-amplifier equipment/device where is conducted the measurement of the phase, which determines bearing. Is possible the application/use also of the mixed methods. According to a phase-amplitude method the first cascade/stages of receiver work in phase mode, then produces the transformation of phase dependences into amplitude, usually using sum-and-difference method. In last/latter stages of receiver-amplifier equipment/device is conducted the amplification with the preservation/retention/maintaining of amplitude dependences. The measurement of bearing at output is conducted on amplitude indicator. In amplitude-phase radio direction finder the construction of receiver-amplifier equipment/device the same as in amplitude, but for the indication of bearing output voltages are converted into the voltages whose phase depends on bearing, and is utilized phase indicator.

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2.4. Errors of radio direction finder.

In §§ 2.2 and 2.3 were examined common/general/total principles and the methods of determining the bearing. In real conditions the work of radio direction finder, is feasible a whole series of the factors, which make the work of radio direction finder worse and calling the errors in the determination of bearing. The analysis of the reasons for these errors and methods of their elimination or at the worst of account during operation represents one of the most important questions of the theory of radio direction finder.

All errors, which are encountered with direction finding, it is possible to divide first of all into errors systematic and random.

Systematic errors cause offset of bearing from true direction in radio station. Determination in this case can be very evenly the direction of bearing it proves to be inaccurate independent of accuracy of reading.

Systematic errors can in principle be removed by the appropriate calibration of direction finder, i.e., by the preliminary determination of its errors from known stations and by the method of the direction finding of the local oscillator, and by the subsequent introduction into the readings of the corresponding corrections. However, the reasons for errors, as we will see further, are so/such numerous and diverse and frequently so they yield with difficulty to account that in practice the fulfillment of this calibration is impossible, and the introduction of all numerous corrections would completely hinder/hamper the use of a radio direction finder even of pi to calibration. In view of this the systematic errors must be as far as possible removed in direction finder. Only the relatively small errors whose regular dependence on very a few factors (one, maximum of two, for example, from azimuth and from frequency) is accurately establish/installed, can be removed by calibration. The part of the systematic errors, of the not removed and not considered in the form corrections, enters in the random errors of radio direction finder.

Random errors are led to the oscillation/vibrations of bearing in time, moreover usually are observed the rapid, slow and very slow oscillations of bearings. With the very large number of observations to one and the same radio station, undertaken for the wide interval of time, average/mean random error is equal to zero, since divergences to one and in the other direction are equiprobable.

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However, in practice the averaging of all oscillations not is always possible. The rapid oscillations of bearings are averaged into the shorter time intervals, than slow ones, and therefore their averaging is realized more frequently. An error in the single observation determines the practical accuracy of radio direction finder. Therefore the random errors of proper to be brought to the possible minimum. Random error is rate/estimated at average/mean quadratic angular error.

Besides the given separation of errors into random and systematic, there is separation according to the character of their calling reasons.

Reasons, the calling errors of radio direction finder, can consist:

1. In operating principle, diagram, the construction of radio direction finder or separate, its parts and in an inaccuracy in the

mounting of external equipment/device. The errors of this kind are caused, therefore, by the properties of instrument itself. We will call them instrument errors. Among them special place occupies the phenomenon, called antenna effect (see § 4.2). The part of the instrument error can be referred to the systematic errors of radio direction finder, another part - to the very slowly changing random error.

2. In effect of heterogeneity of surface along which is propagated electromagnetic wave (coastal effect, effect of remote environment). This component of random error changes very slowly.

3. In phenomana, which occur during propagation of electromagnetic waves (change in plane of polarization, deflection of path of ray/beam from arc of the great circle, arrival to point of reception/procedure of several interfering ray/beams).

As we will see further, the different errors of the propagation of waves cause the rapid and slow oscillations of bearings.

4. In effect of different kind of those who lie near by antenna of system of object/subjects: antennas, wires, metallic masses, trees, structures, hills, etc.

5. In subjectivity of reading of bearing.

Subjective errors depend on a number of factors and first of all on the degree of the indeterminancy/uncertainty of the reading of bearing, for example, on clearness and stability of minimum or maximum of sound in telephone with the auditory reading of bearing, on width and stability of the strip on which is counted off the bearing in automatic visual radio direction finder and, etc. Different observers or one and the same observer, counting off several times bearing to one and the same radio station, they will give the completely not coinciding readings, more or less differing from the true bearing. Magnitude of error in the separate observations the greater, than more the oscillation of bearing, the less determined those sign/criteria on which is counted off the bearing, for example than it is more diffuse, is wider the minimum of audibility with auditory direction finding on the minimum, is wider and more unstable strip with automatic bearing.

Let us examine first subjective errors independent of the reasons, calling them, since the character of these errors does not depend on that, by which precisely reason they are caused. Subsequently let us pass to the examination of other enumerated reasons for errors, determining for each of them value and the character of systematic errors and the degree of their effect on clearness of the reading of bearing.

2.5. Relationship/ratio of the power of signal and interferences at the output of receiver.

Power of interferences at the input of receiver.

The value of the subjective errors to high degree depends on the noise voltage and interferences, which mask signal and which impede the reading of bearing.

For the target/purpose of analysis, it is expedient to divide interferences and noises into two forms: 1) interference and noises with wide continuous spectrum.

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Within the limits of a narrow band of frequencies, passed by receiver, the spectral density of their power can be considered constant, and total power is of equal to the product spectral densities and band [1]; 2) interference with the narrow line spectrum of frequencies.

FOOTNOTE [1]. Here one should understand effective passband. For the majority of selective systems, effective band by approximately 10o/o is more than passband at the level 0.7. ENDFOOTNOTE.

The first form includes space, atmospheric, some man-made interferences and the inherent noise of receiver. To the second - the interference of radio stations. The interference effect of radio stations is examined into § 2.12.

Let us examine the interferences of the first form, which on their origin are divided into external (atmospheric, industrial, etc.) and internal, caused by the fluctuations of current in the antenna, the ducts, the tubes and other cell/elements of receiver.

The level of outside interferences and their spatial distribution strongly depend on time of days and year, latitude of place, frequency and series of other reasons, examine/considered in the courses of radiowave propagation. To evaluate average relationship/ratios, we let us assume that the interferences come evenly from all sides and let us estimate their intensity with a spectral density of the strength of field of A. Value A depends on factors indicated above. The power of outside interferences at the input of receiver comprises [3.1]

$$P_{\text{вп}} = \frac{A^2 h_e^2}{4R_\Sigma} \frac{\eta}{D} B, \tag{2.23}$$

where B is a passband;

h_e -the effective height of antenna;

R_Σ - radiation resistance;

η - efficiency (efficiency);

D - directive gain (derictive gain).

Internal noises, which appear because of the fluctuations of current in antenna with effective resistance $'R_A$, are characterized by spectral power density kT, where k = 1.3•10^{-23} J/deg is Boltzmann constant, T - absolute temperature. The ratio of the total power of internally-produced noise at the input of receiver to the intensity of the noise, caused by only one antenna resistance, he is called factor of noise N. The intensity internal noises is equal to

$$P_{ш}=NkTB. \qquad (2.24)$$

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Inherent noise level at the input of receiver they characterize also by absolute receiver sensitivity - by that strength of applied field with which is created the voltage on input, it is equal to the effective value of noise voltage. From the determination of absolute sensitivity, it follows that

$$\frac{E^2_{a6c}h_e^2}{4R_a}=\frac{E^2_{a6c}h_e^2}{4R_\Sigma}\eta=P_{ш}=NkTB, \quad \text{where} \quad R_a=\frac{R_\Sigma}{\eta}. \qquad (2.24')$$

After using the known relationship/ratios: for symmetrical antenna [3.1]

$$R_\Sigma=\frac{120\pi^2h_e^2}{\lambda^2D},$$

for the grounded antenna

$$R_\Sigma=\frac{240\pi^2h_e^2}{\lambda^2D},$$

we will obtain the expressions of absolute sensitivity for symmetrical and asymmetric antennas respectively:

$$E_{абс}=\sqrt{\frac{480\pi^2NkTB}{\lambda^2D\eta}},\quad E_{абс}=0{,}14\sqrt{\frac{NB\ кгц}{\lambda^2D\eta}},\ \frac{мкв}{м};$$
$$E_{абс}=\sqrt{\frac{960\pi^2NkTB}{\lambda^2D\eta}},\quad E_{абс}=0{,}2\sqrt{\frac{NB\ кгц}{\lambda^2D\eta}},\ \frac{мкв}{м}. \qquad (2.25)$$

[мкв = μV]

The total power of external and internally-produced noise will be equal to

$$P_{пом}=P_{вн}+P_{ш}.$$

Let us find the strength of the applied field, which creates the same power as total power $P_{пом}$. Let us call/name it the noise field intensity

$$E_{пом}=\sqrt{\frac{P_{пом}4R_{\Sigma}}{h_e^2\eta}}=\sqrt{\frac{A^2B}{D}+E^2_{абс}}. \qquad (2.26)$$

The noise field intensity decreases during decrease in N and A, with an increase derictive gain and efficiency, and also with the contraction of the passband of receiver. The question concerning the advisable selection of passband is examined into § 2.6.

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Absolute sensitivity can be improved (i.e. value $E_{абс}$ is decreased) with an increase derictive gain (D) and efficiency(η)

antenna and during a decrease in the band (B) and in the factor of the noise (N).

Calculation of these values is given in chapter 7.

From formula (2.26) it follows that a decrease in value $E_{абс}$ has little effect on that which results intensity of the field of interferences, if by means of selection N and η achieved/reached considerable excess of outside interferences above its own, i.e., if

$$\frac{A^2B}{D} \gg E^2_{абс}.$$

In the examination of this relationship/ratio, one should consider the minimum level of outside interferences, which can be met under the actual conditions of the work of radio direction finder.

The relationship/ratio of the intensity of signal and intensity of noise at the output of the receptor depends on those transformations, by which undergo the signal and noise. In the examination of these transformations, one should distinguish the linear and nonlinear cell/elements of the receiver.

Relationship/ratio of the power of signal and interferences at the output of the linear part of the receiver.

The linear cell/elements of receiver include the feeiers, which filter circuits, amplifier stages, and also frequency converters and synchronous detectors. The usual detectors, which isolate the modulating voltage of signal, are nonlinear.

After the passage of linear cascade/stages, the ratio of the power of signal to the intensity of noises on output is equal to the same relation at the input:

$$\frac{P_{\text{с вых}}}{P_{\text{ш вых}}}=\frac{P_{\text{с}}}{P_{\text{ш}}}=\frac{P_{\text{с}}}{P_{1\text{ш}}B}, \qquad (2.27)$$

where B is the resulting passband of all linear cascade/stages; $P_{1\text{ш}}$ the spectral noise density at input.

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If separate cascade/stages have a band, B_1, B_2 and so forth, then the resulting passband approximately can be found by the formula

$$B=\frac{1}{\sqrt{\frac{1}{B_1^2}+\frac{1}{B_2^2}+\frac{1}{B_3^2}+\dots}},$$

from which evident that the resulting passband is determined in essence by the narrow-band cell/element.

The intensity of noise at output does not depend on the distribution of selectivity between the cascade/stages of receiver in its linear part (high-frequency amplifier, the IF amplifier and so forth). However, during the incorrect distribution of selectivity, is possible the overloading of cascade/stages by noises or the interferences of other stations, the disturbance/breakdown of linearity, the onset of combination interferences and bearing errors. These questions are illuminated in the courses of radio receiving equipment. The effect of powerful interferences on the errors of two-channel radio direction finders is examined into § 8.3.

In formula (2.27) are not taken into account noise sources, available in receiver after its input: tube and the semiconductor devices of the subsequent cascade/stages. The amplification of the first cascade/stage must be selected so that the noise voltages of the second and of all subsequent cascade/stages would be negligible in comparison with the intensive noises of the first cascade/stage.

Relationship/ratio of the intensity of signal and noise at the output of detector.

In nonlinear cell/elements it is necessary to examine together the passage of signal and noise. This question was the object/subject of the number of special investigations [2.2, 2.3], to which one should turn for a comprehensive study.

Will be here given only the short approximate results, in the majority of cases sufficient to evaluate works of direction-finding systems.

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Let us designate the effective value of the voltage of signal carrier frequency on the input of detector U_0 and the effective value of noise voltage $U_{\text{ш}}$. If the passband of the frequencies of that part of the receiver, which precedes detector, $B_{\text{в}}$, and spectral noise density (on voltage) $\beta_{\text{ш}}$, then

$$U_{\text{ш}} = \beta_{\text{ш}} \sqrt{B_{\text{в}}}.$$

We examine the signal, modulated in amplitude, with the depth of modulation M. At the output of detector, we will obtain stress component modulation frequency, caused by signal with effective value $U_{\text{c вых}}$ and the spectrum of the noise voltages, caused by the interaction of the components of noise between themselves and with signal. The effective value of the noise voltage within the limits of the passband $B_{\text{н}}$ of the filter, following after detector, let us designate $U_{\text{ш вых}}$. The ratio of the voltage of signal to noise voltage on the output of the square law detector is determined by the formula

$$\frac{U_{\text{c вых}}}{U_{\text{ш вых}}} = M \frac{U_{\text{c}}}{U_{\text{ш}}} \sqrt{\frac{B_{\text{в}}}{2B_{\text{н}}}} \frac{1}{\sqrt{1 + \frac{M^2}{2} + \frac{U_{\text{ш}}^2}{2U_{\text{c}}^2}\left(1 - \frac{B_{\text{н}}}{2B_{\text{в}}}\right)}}. \qquad (2.28)$$

It is here assumed that $2B_{\text{н}} < B_{\text{в}}$. If $2B_{\text{н}} > B_{\text{в}}$, then into formula one should substitute unity for the relation $2B_{\text{н}}:B_{\text{в}}$. In the majority of cases $2B_{\text{н}} \ll B_{\text{в}}$ and in denominator under radical sign it is possible to disregard $\frac{B_{\text{н}}}{2B_{\text{в}}}$ in comparison with unity. The depth of modulation M is usually less than unity, which makes it possible to disregard the term $M^2/2$.

Formula (2.23) is derived on the assumption that the frequency characteristics of the predetector part of the receiver are rectangular. Investigation shows [2.6] that the form of characteristic with constant passband insignificantly affects the relation of the voltages of signal and noise on output. Therefore to admissibly use formula (2.28) with any form of characteristic, especially because the rectangular characteristic determines the lowest sense of the voltages of signal and noise.

In the case of linear detection of formula, they are obtained more complex.

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However, it is possible to show [2.6] that the results of linear detection differ less than by 7o/o from the results of square-law detection in the identical sense of the voltages of signal and noises on the input of detector. For this reason to admissibly use and for linear detection formula (2.28) or its simplified form:

$$\frac{U_{\text{с вых}}}{U_{\text{ш вых}}}=M\frac{U_{\text{с}}}{U_{\text{ш}}}\sqrt{\frac{B_{\text{в}}}{2B_{\text{н}}}}\,\frac{1}{\sqrt{1+\frac{U_{\text{ш}}^2}{2U_{\text{с}}^2}}}. \qquad (2.29)$$

With powerful signals $U_{\text{с}}>U_{\text{ш}}$ the second term under radical in denominator can be disregarded and then

$$\frac{U_{\text{с вых}}}{U_{\text{ш вых}}}=M\frac{U_{\text{с}}}{U_{\text{ш}}}\sqrt{\frac{B_{\text{в}}}{2B_{\text{н}}}}=M\frac{U_{\text{с}}}{\sigma_{\text{ш}}\sqrt{2B_{\text{н}}}}=M\frac{U_{\text{с}}}{U'_{\text{ш}}}, \quad (2.30)$$

where $U'_{\text{ш}}=\sigma_{\text{ш}}\sqrt{2B_{\text{н}}}$ - is the effective stress of noise on the input in passband $2B_{\text{н}}$.

The relationship/ratio of the voltages of signal and noise is determined with powerful signals exclusively by the passband of low pass filter, in this case it is assumed that its passband already, than the half of the passband of high-pass filter. With very weak signals it is possible in expression (2.29) to disregard unity in comparison with $\frac{U_{ш}^2}{2U_{с}^2}$:

$$\frac{U_{с\,вых}}{U_{ш\,вых}} = \sqrt{2}M\frac{U_{с}^2}{U_{ш}^2}\sqrt{\frac{B_{в}}{2B_{н}}} = \sqrt{2}M\frac{U_{с}^2}{U_{ш}^2}\cdot\frac{1}{\sqrt{B_{в}2B_{н}}}. \quad (2.31)$$

In this case the relationship/ratio of the voltages of signal and noise on output depends to identical degree on passbands on high and in low frequency, this relationship/ratio proves to be inversely proportional to square root of certain equivalent band, equal to geometrical mean from passbands in high and low frequency.

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2.6. Selection of passband.

Maximum passband is determined by that navigational information which will bear the signal. One should distinguish search, either the initial determination of bearing, and tracking, or the consecutive determination of bearing with the motion of the oriented radio station. During search the signal must be reveal/detected in certain direction for time $T_н$, which is assigned by the duration of signal or by the conditions of work. In systems with motionless antennas, the signal is detected independent of its direction and passband must be sufficient only for providing the complete establishment of signal for time $T_н$. It is known that the passband is connected with set-up time by the relationship/ratio

$$B\approx\frac{1}{T_н}, \qquad (2.32)$$

which determines required passband. In systems with the rotatable antennas, which possess directivity, the retention time of signal within the limits of the main lobe of radiation, which has equivalent $\theta_э$, will comprise

$$\tau=\frac{T_н\theta_э}{2\pi} \qquad (2.33)$$

and the passband, which ensures the establishment of signal, will be

$$B\approx\frac{1}{\tau}=\frac{2\pi}{\theta_э T_н}. \qquad (2.34)$$

Here $T_н$ is taken as equal to the period of the rotation of antenna. Actually passband must be wider, since usually single detection insufficient.

The comparison of formulas (2.32) and (2.34) shows that the passband and, consequently, also the intensity of noises, is considerably above in the case of the rotatable directed system.;

however, even, the power of signal, which can be extracted from the incident wave in this case, is above, since it is proportional to k. n. d. Since here is examined the effect of directivity only in horizontal plane,

$$D_r = \frac{2\pi}{\theta_0}.$$

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The ratio of the power of signal to the intensity of noise will be proportional

$$\frac{D_r}{B} \approx T_n$$

for both systems.

With tracking the moved target/purpose, the rate of the establishment of processes in receiving indicator must be sufficient in order to reflect changes in the bearing. If are assigned the permissible error of measurement Δ and angular rate dθ/dt, then time of the determination of bearing in any position of object must not exceed Δ/(dθ/dt) and, therefore, passband must be not less than

$$B \approx \frac{1}{\Delta}\frac{d\theta}{dt}. \qquad (2.35)$$

A greatest change in the bearing of the object, which moves at a rate of v at a distance R from direction finder, will be when rate is perpendicular to line of bearing. In this case

$$\frac{d\theta}{dt} \approx \frac{v}{R}$$

and formula (2.35) can be rewritten in the form

$$B \approx \frac{1}{\Delta}\frac{v}{R}.$$

Substitution the given formula of the most unfavorable values of the entering it quantities shows that on the basis of the rate of processing about bearing is admissible the application/use of narrow passbands of the order of the units of hertz.

The virtually used in radio direction finders passbands frequently are considerably wider, which is explained by the technical difficulties of the fulfillment of radio direction finders with very narrow passband. These difficulties first of all are caused by swinging of heterodynes and filtering circuits of receiving indicator, and also the generator of transmitter.

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Swinging leads first of all to the need for frequency search with tuning within the limits of certain band of frequencies $F_{\text{н}}$ besides search examined above azimuth. In the theory of frequency search, is known relationship/ratio [2.1]

$$B \approx \sqrt{\frac{F_{\text{н}}}{\tau}},$$

where τ is time of observation.

Utilizing (2.33), we obtain

$$B \approx \sqrt{\frac{F_{\text{н}}}{T_{\text{н}}}\frac{2\pi}{\vartheta_{1}}}. \qquad (2.36)$$

The band, determined by formula (2.36), can be considerably wider than determined by formula (2.34).

In some systems of radio direction finders (two-channel, phase-meter, etc.) the detuning of receiving indicator relative to signal frequency or the detuning of its cell/elements of relatively each other leads to the onset of the errors whose value grow/rises (other conditions being equal,) with the contraction of passband. This fact also forces to apply passbands wider, than maximum permissible. These questions are examined in more detail in chapter 8.

In the case of application/use at the input of the receiving indicator of local modulation, wide passband is required only in the cascade/stages of the high and intermediate of frequencies (to detector). Passband in low frequency (after detector) can be undertaken narrow in accordance with asymptotic relations (2.34), (2.36). During the interference level of this circuit, one should consider the effect of detection on the relationship/ratio intensity of signal and noise - with weak signals unfavorable (see § 2.5).

In certain cases for simplification in the circuit of equipment/device, it is desirable to realize reception of supplementary information (identification signals, official calls etc.) on the same receiving channel with the aid of which is fulfilled the direction finding.

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The passband, required for this (for example, with telephony), can be considerably wider than with direction finding.

2.7. Sensitivity with the reading of bearing on the minimum for audition.

When the output voltage of signal during the rotation of antenna falls below known value, operator's eye/ear ceases to accept signal as a result of the masking effect of noises. Let this disappearance of audibility occur within the limits of angle from $\theta - \theta_0$ to $\theta + \theta_0$, where θ is the true bearing. The angle $2\theta_0$ within which is not distinguished the audibility in telephone during the rotation of antenna, he is called the angle of silence [1].

FOOTNOTE [1]. The subsequent conclusion/derivations are valid in the case of antenna effect when audibility with direction finding completely does not disappear, but is observed the angle of the equal to audibility. ENDFOOTNOTE.

Determining bearing for audition, we can establish/install antenna in any position within this angle. Consequently, a maximum error in the determination of bearing is equal to θ_0. The reading of bearing somewhat is more precisely formulated, if we find the boundaries of the angle of silence and then to accept for bearing their average value. As a result of the indeterminancy/uncertainty of the boundaries of the angle of silence, the possibility of error is retained also in this case. An average (in absolute value) error in this determination can be accepted on the basis of experimental data as the equal from one eighth to one fourth of the angle of silence depending on the skills of operator and time, spent on the fulfillment of reading.

Let us find the dependence of the angle of silence on noise level at the output of radio direction finder. In this case, by account the physiological special feature/peculiarities of hearing aid. Audibility is proportional to the logarithm of the power of sound. If P_0 is the power, which corresponds to threshold of audibility, then audibility will be (in dB)

$$L = 10\lg\frac{P}{P_0}.$$

In radio direction finders they utilize beat reception with the aid of the heterodyne, voltage on detector from which considerably exceeds as voltage of signal, so, and noise voltage. The process of detection (frequency conversion) under this condition it is possible to consider linear and the relation of the intensities of signal and noises at output equal to the same relation at input.

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Net power, developed with signal at the input of receiver, is equal to

$$P_{\text{c}} = \frac{E^2 h_e^2}{4R_\Sigma}\eta_1 F^2(\theta),$$

where F(θ) - the standardized/normalized radiation pattern in horizontal plane.

The power of interferences was determined by formulas (2.23) and (2.24'). The total power of signal and interferences is equal to

$$P = P_{\text{c}} + P_{\text{пом}} = [E^2F^2(\theta) + E^2_{\text{пом}}]\,\frac{h_e^2}{4R_\Sigma}\eta_1.$$

Let us select the beginning of the calculation of the angles of rotation of antenna so that with $\theta = 0$ $F(\theta) = 0$. When antenna is located accurately in initial position $\theta = 0$, audibility is equal to

$$L_1 = 10\lg\frac{E^2_{\text{пом}} h_e^2 \eta}{4R_\Sigma P_e}.$$

During the rotation of antenna through small angle θ_0, the audibility will become equal to

$$L_2 = 10\lg\frac{E^2_{пом} + E^2F^2(\theta_0)}{4R_\Sigma P_0}h_д^2\eta.$$

We hence obtain an increment in the audibility

$$\Delta L = L_2 - L_1 = 10\lg\left[1 + \frac{E^2F^2(\theta_0)}{E^2_{пом}}\right].$$

Considering in last/latter expression the second term under log sign small, approximately we obtain

$$\Delta L = 4{,}34\frac{E^2F^2(\theta_0)}{E^2_{пом}}$$

or

$$E = E_р = \sqrt{\frac{\Delta L}{4{,}34}}\frac{E_{пом}}{F(\theta_0)}. \qquad (2.37)$$

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The obtained strength of field $E_р$ determines the real sensitivity of radio direction finder, i.e., that strength of field which is required for the reading of bearing with the error, which does not exceed the assigned magnitude.

Being given any increment in the audibility ΔL during rotation by the antenna of system through angle θ_0 from the position of bearing, it is possible from formula (2.37) to calculate the required strength of field.

For determining the instrument/tool sensitivity of radio direction finder, i.e., sensitivity in the absence of outside interferences, it suffices to place A = 0. In this case, $E_{\text{ном}} = E_{\text{абс}}$ and the instrument/tool sensitivity

$$E_{\text{ин}} = \sqrt{\frac{\Delta L}{4,34} \frac{E_{\text{абс}}}{F(\theta_0)}}. \qquad (2.38)$$

For the small values of angle θ_0, only and being of interest in practice, with direction finding on the minimum, taking into account that F (0) = 0, the value of function $F(\theta_0)$ can be presented approximately in the form

$$F(\theta_0) = F'(0)\,\theta_0.$$

After substituting last/latter expression into formulas (2.37) and (2.38), we will obtain

$$E_{\text{р}} = \sqrt{\frac{\Delta L}{4,34} \frac{E_{\text{ном}}}{\theta_0 F'(0)}}, \quad E_{\text{ин}} = \sqrt{\frac{\Delta L}{4,34} \frac{E_{\text{абс}}}{\theta_0 F'(0)}}. \qquad (2.39)$$

Hearth ΔL is here understood the minimum increment in the audibility, detected by eye/ear. This value depends on the subjective special feature/peculiarities of observer, passband of receiver, absolute sound intensity, pitch of the tone and other factors.

Experimental data show that at average value ΔL lie/rests within limits 0.5-1 dB.

The angle θ_0 on boundary of which the signal is detected that during the rotation of antenna, it is equal to one-half angle of silence and determines the accuracy of reading of bearing.

Let us accept as an example the permissible reading error 0.5°.

Since a reading error comprises (0.12-0.25)$2\theta_0$, the angle of silence must be $2\theta_0$ = 2-4°.

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Under these assumptions from (2.39) the real sensitivity of radio direction finder will be

$$E_{\text{р}} = (10 \div 27{,}5)\frac{E_{\text{пом}}}{F'(0)},$$

a the instrument/tool sensitivity

$$E_{\text{ин}} = (10 \div 27{,}5)\frac{E_{\text{абс}}}{F'(0)}.$$

From this example it is evident that the instrument/tool and real sensitivity with the assigned reading error oscillates within sufficiently wide limits depending on observer's subjective qualities. For the possibility of the objective comparison of radio direction finders, it is possible to standardize ΔL and θ_0 with the assigned accuracy of reading of bearing. Then instrument/tool and real sensitivities ($E_{\text{ин}}$ and $E_{\text{р}}$) become completely determined, since absolute sensitivity, the noise field intensity and radiation pattern are determined experimentally objectively.

From expressions (2.39) it follows that the product of the angle of silence by the strength of field is a constant value for this direction finder and the assigned wavelength:

$$M_{\text{р}} = 2\theta_0 E_{\text{р}} = 2\sqrt{\frac{\Delta L}{4{,}34}\frac{E_{\text{пом}}}{F'(0)}},$$

$$M_{\text{ин}} = 2\theta_0 E_{\text{ин}} = 2\sqrt{\frac{\Delta L}{4{,}34}\frac{E_{\text{абс}}}{F'(0)}}.$$

Accepting ΔL = 0.5-1 dB and expressing angle θ_0 in degrees, we obtain

$$M_{\text{р}}=(35\div 55)\frac{E_{\text{пом}}}{F(0)},\quad M_{\text{пп}}=(35\div 55)\frac{E_{\text{анс}}}{F(0)}.$$

The product of the angle of silence by the strength of field calls respectively the real or instrument/tool module/modulus of the sensitivity of radio direction finder and frequently they accept as the measure of its sensitivity.

During the experimental determination of the module/modulus of sensitivity, finds the value of the angle of silence at the measured strength of the field of signal. The definition of the angle of silence, as has already been indicated that strongly it depends on observer's subjective special feature/peculiarities, what is an essential deficiency/lack in the application/use of a module/modulus of sensitivity as the measure of the evaluation of the sensitivity of direction finder.

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Let us use the given formulas to radio direction finder with sinusoidal radiation pattern (loop antenna, two spaced vertical wire antennas, the goniometric system).

Radiation pattern and its first-order derivative will be

$$F(\theta) = \sin\theta, \quad F'(\theta) = \cos\theta,$$

with $\theta = 0$

$$F'(0) = 1.$$

Then for sensitivity and for the module/modulus of sensitivity we obtain the expressions:

$$\begin{aligned} E_{р} &= (10 \div 27{,}5)\, E_{ном}, \\ E_{пн} &= (10 \div 27{,}5)\, E_{абс}, \\ M_{р} &= (35 \div 55)\, E_{ном}, \\ M_{пн} &= (35 \div 55)\, E_{абс}. \end{aligned}$$

2.8. Sensitivity of radio direction finder during the use of comparison method.

With direction finding for audition according to comparison method, or as it occasionally referred to as, according to equisignal method, they revolve antenna system, simultaneously changing over any cell/elements, which change radiation pattern (for example, see Fig. 2.7).

Bearing is read in that antenna position of system, in which the audibility does not change during switchings.

Figures 2.15 depicts to the dependence of voltage U_T, removed with the anten of system, from angle θ in two positions of 1 and of 2 switches.

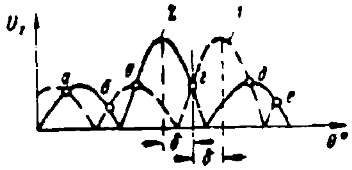

Fig. 2.15. Displaced radiation patterns, utilized with comparison method.

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The positions of switch correspond to the displacement of radiation patterns to angle δ.

Letters a, b, c, d, e, f designated the points at which is satisfied the condition of the equal to audibility. As can be seen from figure, there are many such points, which leads to the indeterminancy/uncertainty of bearing. For the exception/elimination of many readings, it is necessary to ensure single-lobe radiation pattern or at least to considerably attenuate/weaken all minor lobes.

After selecting as reference point the angle, which corresponds to point d of the intersection of the main lobes of radiation, we can present the first diagram as $F(\theta + \delta)$, and the second as $F(\theta - \delta)$. Considering both diagrams symmetrical relative to straight lines $+\delta$ and $-\delta$ respectively, we can write

$$|F(\delta)| = |F(-\delta)|.$$

Due to inaccuracy in the determination of the equality of audibility near point d, can be obtained an error in the bearing. For example, the equality of audibility can be determined not at angle $\theta = 0$, but at certain angle $\theta = \theta_0$. Magnitude of error depends on the relationship/ratio of the voltages of signal and interferences on the output of radio direction finder.

The sensitivity of radio direction finder with direction finding according to comparison method let us call/name, as earlier, that minimum strength of field, which provides the possibility of direction finding with the error, which does not exceed certain assigned magnitude $\Delta\theta$.

For determining sensitivity, let us find the total power of signal and interferences on the output of radio direction finder in two positions of switch, which correspond to radiation patterns 1 and 2:

$$P_1 = P_{\text{пом}} + E^2 F^2 (\theta_0 + \delta) \frac{h_{\text{д}}^2 \eta}{4 R_\Sigma},$$

$$P_2 = P_{\text{пом}} + E^2 F^2 (\theta_0 - \delta) \frac{h_{\text{д}}^2 \eta}{4 R_\Sigma},$$

where $P_{\text{пом}}$ it is determined by formula (2.26).

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An increment in the audibility during transition from position of 1 to position of 2 will be

$$\Delta L = 10\lg\frac{P_1}{P_2} = 10\lg\frac{E^2_{\text{пом}} + E^2F^2(\theta_0+\delta)}{E^2_{\text{пом}} + E^2F^2(\theta_0-\delta)}. \qquad (2.40)$$

It is decomposed expression for a radiation pattern in Taylor series, considering angle δ small:

$$\begin{aligned}F(\theta_0+\delta) &= F(\delta) + \theta_0 F'(\delta),\\ F(\theta_0-\delta) &= F(\delta) - \theta_0 F'(\delta),\\ F^2(\theta_0+\delta) &= F^2(\delta) + 2F(\delta)F'(\delta)\theta_0,\\ F^2(\theta_0-\delta) &= F^2(\delta) - 2F(\delta)F'(\delta)\theta_0.\end{aligned}$$

After substituting these expressions into equation (2.40) and after using the formula of the approximation calculus

$$\lg\frac{1+x}{1-x} = 0{,}868x,$$

when x << 1, we will obtain

$$\Delta L = 8{,}68\,\frac{2\theta_0 E^2F(\delta)F'(\delta)}{E^2_{\text{пом}} + E^2F^2(\delta)}.$$

Hence we find

$$E_{\text{р}} = \sqrt{\frac{\Delta L}{17{,}36\theta_0 F(\delta)F'(\delta) - \Delta L F^2(\delta)}}\,E_{\text{пом}}. \qquad (2.41)$$

Last/latter expression determines the real sensitivity of radio direction finder.

For determining instrument/tool sensitivity, let us place A = 0. Then we obtain

$$E_{\text{нп}} = \sqrt{\frac{\Delta L}{17{,}36\theta_0 F(\delta)F'(\delta) - \Delta L F^2(\delta)}}\,E_{\text{абс}}. \qquad (2.42)$$

From the obtained expressions it is evident that the sensitivity depends on the value of the angle δ of the displacement of diagram. There is a optimum angle of shift of diagrams, which ensures the best sensitivity.

As an example of the application/use of the given formulas, let us examine the circuit, presented in Fig. 2.7.

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The radiation pattern of this system is determined by the formula

$$F(\theta \pm \delta) = \frac{\cos\theta \pm n\sin\theta}{\sqrt{1+n^2}}, \qquad (2.43)$$

where $n = \frac{h_B}{h_A}$ — is relation of the effective height of framework B and A. Signs ± in formula correspond to two positions of switch D. From expression (2.43) it is possible to obtain

$$F(\theta \pm \delta) = \sin(\theta \pm \delta), \quad F'(\theta \pm \delta) = \cos(\theta \pm \delta),$$

where

$$\cos\delta = n\sin\delta, \quad \operatorname{tg}\delta = \frac{1}{n},$$

$$\sin\delta = \frac{1}{\sqrt{1+n^2}}, \quad \cos\delta = \frac{n}{\sqrt{1+n^2}}.$$

After substituting these expressions with $\theta = 0$ into formulas (2.41) and (2.42), we find the real and instrument/tool sensitivity

$$E_{\text{р}} = \sqrt{\frac{\Delta L}{8.68\theta_0 \sin 2\delta - \Delta L \sin^2\delta}}\, E_{\text{ном}}, \qquad (2.44)$$

$$E_{\text{нн}} = \sqrt{\frac{\Delta L}{8,68\theta_0 \sin 2\delta - \Delta L \sin^2\delta}}\, E_{\text{абс}}. \qquad (2.45)$$

The strength of field $E_{\text{р}}$ and $E_{\text{нн}}$ have a minimum with or with

$$\left.\begin{aligned} \operatorname{tg} 2\delta_{\text{опт}} &= \frac{2n}{n^2-1} = \frac{17{,}36\theta_0}{\Delta L} \\ n_{\text{опт}} &= \frac{\Delta L + \sqrt{(\Delta L)^2 + (17{,}36\theta_0)^2}}{17{,}36\theta_0}. \end{aligned}\right\} \qquad (2.46)$$

The minimum strength of field, which corresponds to the best real sensitivity, will be, considering ΔL = 1 dB:

$$E_{p\,\text{мин}} = E_{\text{ном}} \sqrt{\frac{2}{\sqrt{1+(17{,}36\theta_0)^2}-1}}. \qquad (2.47)$$

Respectively for minimum instrument/tool sensitivity it is possible to write

$$E_{\text{пн.мин}} = E_{\text{абс}} \sqrt{\frac{2}{\sqrt{1+(17{,}36\theta_0)^2}-1}}. \qquad (2.48)$$

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In Fig. 2.15 is constructed the dependence $E_p/E_{\text{ном}}$ on angle equal to audibility $2\theta_0$ at different values of n and with ΔL = 1 dB. In this same figure is represented the same dependence $\frac{E_p}{E_{\text{ном}}}$ on $2\theta_0$ with direction finding on the minimum.

From the figure one can see that the sensitivity with minimum direction-finding method proves to be above $\left(\frac{E_p}{E_{\text{ном}}}\right.$ less) at small angles of the equal to audibility. At the large angles of the equal to audibility, on the contrary, is more favorable the comparison method. This same conclusion/derivation can be otherwise expressed as follows: in the large ratio of the strength of the field of signal to the noise field intensity for direction finding, is more favorable minimum method, in small ratio of the strength of the field of signal to the noise field intensity - comparison method.

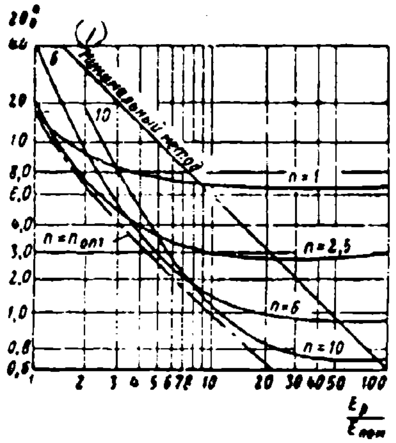

Fig. 2.16. Dependence of angle of equal to audibility on interference level.

Key: (1). Minimum method.

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From the examination of formulas (2.39), the relating to direction finding on the minimum, and formulas (2.41) and (2.42) for an equisignal method follows that in both cases with the assigned accuracy of reading (Δθ) the sensitivity of radio direction finder depends on the following factors: the sensitivity of recorder (the hearing aid of operator), the slope/transconductance of radiation pattern near the value of angle, which corresponds to the minimum of reception, and from interference level. In the case of equisignal method, has a value also the amount of the mutual displacement of radiation patterns. The effect of last/latter factor was already examined.

The sensitivity of radio direction finder is improved with an increase in the slope/transconductance of radiation pattern $F'(\theta)$, i.e., during the application/use of the highly directional antenna systems. A deficiency/lack in such antenna systems is an which is inherent in them multitude of directions of zero reception, which leads to indeterminancy/uncertainty with the reading of bearing. This indeterminancy/uncertainty can be removed by finding the main lobes of radiation of the sign/criterion of the greatest intensity of reception.

On short and especially on medium-frequency waves the execution of the rotatable highly directional antenna systems is extremely hinder/hampered also in certain cases virtually impracticably, since the size/dimensions of antennas with acute/sharp radiation pattern must be great in comparison with wavelength.

The use of the pencil-beam antennas in usual goniometric system is impossible. In § 3.11 are given the descriptions of the methods of use by a motionless antenna of system with acute/sharp radiation pattern for direction finding. With the shortening of wavelength, the size/dimensions of antenna systems decrease also on ultra short waves, especially into the ranges of microwaves, highly directional the rotatable direction-finding antennas prove to be easily feasible.

For lowering in the interference level both external and internal the passband of the frequencies of the receiver must be made narrowest possible. It is necessary to keep in mind that the effective bandwidth, entering formulas (2.39) and (2.41), must consider the selectivity of the hearing aid of observer. The character of noise in telephone with the contraction of passband changes, approaching the simple tone, which coincides with the tone of signal, which impedes the discrimination of signal. For the indicated reason the passband of auditory direction finder can be accepted by the approximately equal to 100 Hz independent of the passband of radio engineering circuit.

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The latter nevertheless must be made narrowest possible for a decrease in the interferences from radio stations.

The ways of a decrease in the resulting noise field intensity are examined into § 2.5.

2.9. Sensitivity with direction finding on the minimum of the depth of modulation.

Modulation of signal whose depth depends on bearing, is realized by the addition of the modulated voltage, which enters from the directional antenna, for example the framework, with voltage from the omnidirectional antenna, see Fig. 2.9). The block diagram of radio direction finder is represented in Fig. 2.17. If E is strength of field, k_p — the effectiveness of the framework, k_a — the effectiveness of antenna, k_{py} and k_{ay} — the transmission gains of a modulation-amplifier equipment/device of framework and amplifier of antenna respectively, Ω/2π - modulation frequency, ω/2π - signal frequency, then at the point of the addition of the voltages (at the input of high-frequency amplifier) we will obtain voltage from the framework

$$u_p = \sqrt{2} E k_p k_{py} F(\theta) \sin \Omega t \sin \omega t$$

and voltage from the antenna

$$u_a = \sqrt{2} E k_a k_{ay} \sin \omega t.$$

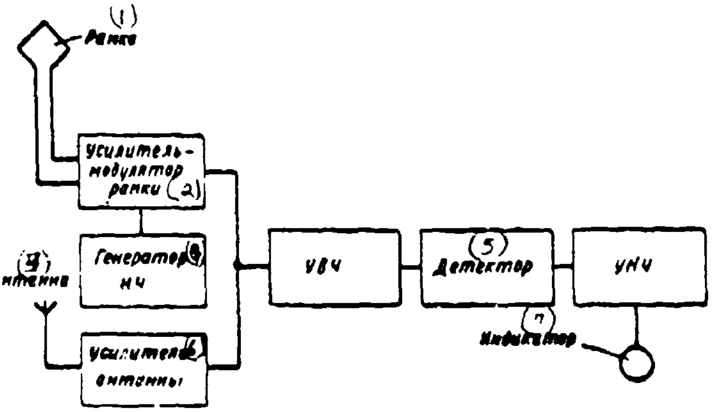

Fig. 2.17. Block diagram of radio direction finder for direction finding on the minimum of the depth of modulation.

Key: (1). Framework. (2). Amplifier-modulator of the framework. (3). Antenna. (4). Generator of L.F. (5). Detector. (6). Amplifier of antenna. (7). Indicator.

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The total voltage

$$\begin{aligned} u &= u_p + u_a = \\ &= \sqrt{2} E k_a k_{ay}\left[1 + \frac{k_p k_{py}}{k_a k_{ay}} F(\theta)\sin\Omega t\right]\sin\omega t = \\ &= \sqrt{2} U_c [1 + M\sin\Omega t]\sin\omega t \qquad (2.49) \end{aligned}$$

is the modulated oscillation/vibration. The depth of modulation, which is the function of bearing, is equal to

$$M = \frac{k_p k_{py}}{k_a k_{ay}} F(\theta) = M_0 F(\theta).$$

Along with signal at the input of a modulation-amplifier equipment/device of framework and amplifier of antenna, will operate noises and interferences the spectral density (on voltage) $\mathcal{E}_{\text{пр}}$ and $\mathcal{E}_{\text{па}}$ respectively:

$$\left.\begin{aligned}\mathcal{E}_{\text{пр}}&=\sqrt{\mathcal{E}^2_{\text{шр}}+\frac{A^2k_p^2}{D_p}},\\ \mathcal{E}_{\text{па}}&=\sqrt{\mathcal{E}^2_{\text{ша}}+\frac{A^2k_a^2}{D_a}},\end{aligned}\right\}\qquad(2.50)$$

where $\mathcal{E}_{\text{шр}}$ — the spectral density of inherent noise at the input of a modulation-amplifier equipment/device of the framework;

$\mathcal{E}_{\text{ша}}$ — the spectral density of inherent noise at the input of the amplifier of antenna;

A - the spectral density (on the strength of field) of outside interferences.

Inherent noise in the channels of the framework and antennas originate from different sources and are statistically independent. Outside interferences with certain approach/approximation let us examine just as statistically independent variables, in view of the fact that directional characteristic of the framework and antenna completely different.

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The resulting noise voltage and interferences on the input of high-frequency amplifier will be equal

$$U_{\text{пом}} = \sqrt{(\mathcal{E}^2_{\text{шр}} k^2_{\text{ру}} + \mathcal{E}^2_{\text{па}} k^2_{\text{ау}}) B_{\text{в}}} = k_{\text{а}} k_{\text{ау}} \sqrt{\left[\left(\frac{\mathcal{E}^2_{\text{шр}}}{k^2_{\text{р}}} + \frac{A^2}{D_{\text{р}}}\right) M^2 + \frac{\mathcal{E}^2_{\text{ша}}}{k^2_{\text{а}}} + \frac{A^2}{D_{\text{а}}}\right] B_{\text{в}}}. \quad (2.51)$$

where $B_{\text{в}}$— the effective (is noise) passband of high-frequency amplifier.

The relationship/ratio of the voltage of signal and noises, determined by formulas (2.49), (2.51), will be preserved at the output of linear high-frequency amplifier, i.e., at the input of detector.

For determining the relation of the voltages of signal and noises on the output of detector, we will use formula (2.29).

$$\frac{U_{\text{с вых}}}{U_{\text{п вых}}} = M_0 \frac{U_{\text{с}}}{U_{\text{пом}}} \sqrt{\frac{B_{\text{в}}}{2B_{\text{н}}}} \frac{F(\theta)}{\sqrt{1 + \frac{U^2_{\text{пом}}}{2U^2_{\text{c}}}}}.$$

Substitution into this formula of value U_{c} and $U_{\text{пом}}$ according to (2.49) and (2.51) gives

$$\frac{U_{\text{с вых}}}{U_{\text{п вых}}} = M_0 \sqrt{\frac{B_{\text{в}}}{2B_{\text{н}}}} \times$$

$$\times \frac{\sqrt{2} E^2 F(\theta)}{\sqrt{\left(\frac{\mathcal{E}^2_{\text{пр}}}{k^2_{\text{р}}} M_0^2 + \frac{\mathcal{E}^2_{\text{па}}}{k^2_{\text{а}}}\right) B_{\text{в}}} \sqrt{2E^2 + \left(M_0^2 \frac{\mathcal{E}^2_{\text{пр}}}{k^2_{\text{р}}} + \frac{\mathcal{E}^2_{\text{па}}}{k^2_{\text{а}}}\right) B_{\text{в}}}}. \quad (2.52)$$

The low-frequency stress component of noise on output, operating on indicator, they will cause chaotic beat of the pointers whose root-mean-square $\alpha_{\text{ш}}$ value is proportional $U_{\text{п вых}}$. The voltage of signal will cause systematic throw of pointer α, proportional U_{c}. Observer notes the throw of pointer, caused by signal, if it not too little relative to chaotic oscillation/vibrations, i.e., with

$$\frac{\alpha}{\alpha_{\text{ш}}}=\frac{U_{\text{c вых}}}{U_{\text{п вых}}}\geq C,$$

where C - the coefficient of discernability and in experimental data is equal to 0.5-1. For determining sensitivity, i.e., the minimum strength of field, which ensures a sufficient coefficient of discernability during the assigned divergence of the framework θ_0 from the position, which corresponds to the bearing with which F (θ) = 0, is decomposed by F (θ) in Taylor series, being limited to the first, by the term

$$F(\vartheta)=\vartheta_0 F'(\vartheta),$$

where θ_0 is small deviation of angle θ from zero.

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Let us designate

$$\frac{C\sqrt{\frac{2B_{\text{н}}}{B_{\text{э}}}}}{\sqrt{2}\,\theta_0 F'(0)}=C_0 \qquad (2.53)$$

and is solved equation (2.52) relative to E

$$E=\sqrt{\frac{\left(\frac{G^2_{\text{нр}}}{k^2_{\text{р}}}M_0^2+\frac{G^2_{\text{нз}}}{k^2_{\text{н}}}\right)B_{\text{э}}\left(C_0^2+\sqrt{C_0^4+C_0^2M_0^2}\right)}{M_0^4}} \qquad (2.54)$$

Let us examine the effect of different factors on sensitivity. A decrease in the first factor under radical sign can be reached by an increase in the effectiveness of the framework and antenna. Copying this factor in the expanded/scanned form, we obtain

$$\frac{\mathcal{E}^2_{\text{пр}}}{k^2_p}M_0^2+\frac{\mathcal{E}^2_{\text{па}}}{k^2_a}=\left(\frac{\mathcal{E}^2_{\text{шр}}}{k^2_p}+\frac{A^2}{D_p}\right)M_0^2+\frac{\mathcal{E}^2_{\text{ша}}}{k^2_a}+\frac{A^2}{D_a}.$$

An increase k_p and k_a is expedient until the voltages of the inherent noise of the channels of the framework and antenna, converted into the circuit of the framework and antenna, become small in comparison with outside interferences. This conclusion/derivation coincides with analogous conclusion/derivation for direction finding in the minimum. With the predominance of outside interferences, it is possible to consider that

$$\frac{\mathcal{E}^2_{\text{пр}}}{k^2_p}\approx\frac{\mathcal{E}^2_{\text{па}}}{k^2_a}.$$

Formula (2.54) will take the form

$$E=\frac{\mathcal{E}_{\text{пр}}\sqrt{B_0}}{k_p}C_0\sqrt{\frac{(1+M_0^2)\left(1+\sqrt{1+(M_0C_0)^2}\right)}{M_0^2}}. \quad (2.54')$$

Figures 2.18 depicts the dependence of sensitivity on the modulation factor at the different values of parameter C_0. The required strength of field sharply descends with an increase in the modulation factor up to $M_0 = 1$.

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With the further increase M_0, is observed either minimum E at the low values C_0 or the slow, smooth decrease E at the large values C_0. Hence it follows that is expedient to select M_0 equal to 1-1.5. The large values M_0, having little effect on sensitivity, are unfavorable in other respects (see § 8.1).

With that which was assigned M_0 the sensitivity is improved during a decrease in parameter C_0. From formula (2.53) it follows that for decrease C_0 it is necessary to utilize the narrowest possible frequency band in low frequency, and also, if this is possible, to increase mutual conductance of directivity. Usually in this type radio direction finders $F(\theta) = \sin\theta$ and $F'(0) = 1$.

The given results, obtained on the assumption that predominate outside interferences, retain their value also for that case when predominate inherent noise.

If we accept in accordance with the given considerations $M_0 = 1$, $C_0 = 0.25-1$ and $C = 0.5$, then real sensitivity will be expressed by the formula, obtained from expression (2.54'):

$$E_р = \frac{\mathcal{E}_{пр}}{k_р}(1,5 \div 2,2)\frac{\sqrt{2B_н}}{2\theta_0 F'(0)} =$$

$$= (90 \div 130)\frac{\mathcal{E}_{пр}}{k_р}\frac{\sqrt{2B_н}}{2\theta^{\circ}_0 F'(0)}.$$

Passband $B_н$ is determined by all circuit, following after detector, including the band of the susceptibility of indicator. The latter quite frequently determines the resulting passband.

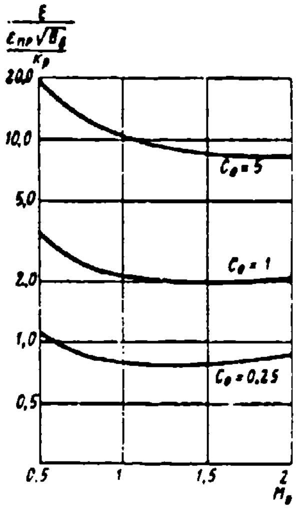

Fig. 2.18. Dependence of sensitivity on modulation factor.

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2.10. Sensitivity with direction finding by phase-difference method.

The measurement of phase consists always of the comparison of two oscillation/vibrations a phase difference of which is determined. It is possible to distinguish two cases: 1) one of the oscillation/vibrations it is virtually free from interferences and only the second is accompanied by interferences, 2) both oscillation/vibrations contain both the signal and the interferences.

Let us examine the first case. It is known that noise voltage can be presented as oscillation/vibration with random amplitudes U_{noi} and random phases ψ_{noi}. Density distribution of the probability of amplitudes follows Rayleigh's law, and phases are accepted equiprobable. Store/adding up the voltage of signal and disturbing voltage, we obtain the resulting voltage with phase displacement ψ relative to signal (Fig. 2.19). It is obvious that value ψ is also random variable. Probability density of its is determined by formula [2.2]

$$W(\psi)=\frac{1}{2\pi}e^{-q^2}+\frac{q\cos\psi}{\sqrt{\pi}}F(\sqrt{2}q\cos\psi)e^{-q^2\sin^2\psi},$$

where q is ratio of actual stress of signal to the effective value of disturbing voltage; $F(\sqrt{2}q\cos\psi)e^{-q^2\sin^2\psi}$ — Laplace function.

With weak signals (q $\ll$ 1), expanding Laplace function in a series, it is possible to obtain

$$W(\psi)=\frac{1}{2\pi}+\frac{q\cos\psi}{2\sqrt{\pi}}.$$

Average value ψ is equal to zero, and the root-mean-square deviation

$$\sigma_1=\sqrt{\frac{\pi^2}{3}-2\sqrt{\pi}q}. \qquad (2.55)$$

With powerful signals (q $\gg$ 1), utilizing an asymptotic representation of Laplace function, we obtain

$$W(\psi)=\frac{q}{\sqrt{\pi}}e^{-q^2\psi^2}.$$

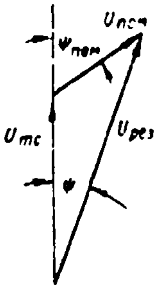

Fig. 2.19. Addition of voltages of signal and noise.

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Distribution normal relatively ψ. with mean deviation $\psi_{cp} = 0$ and with the root-mean-square deviation

$$\sigma_1 = \frac{1}{\sqrt{2q}}. \qquad (2.56)$$

The dependence of root-mean-square deviation σ_1 from q with respect to formulas (2.55) and (2.56) is represented in Fig. 2.23. The section between the fields of the applicability of both formulas is interpolated. In the case in question the obtained deviations of phase under the effect of noise voltage are full deflections, since the phase of the second oscillation is rigid.

In the second case the phase of the second oscillation is also subjected to divergence. The resulting divergence grow/rises.

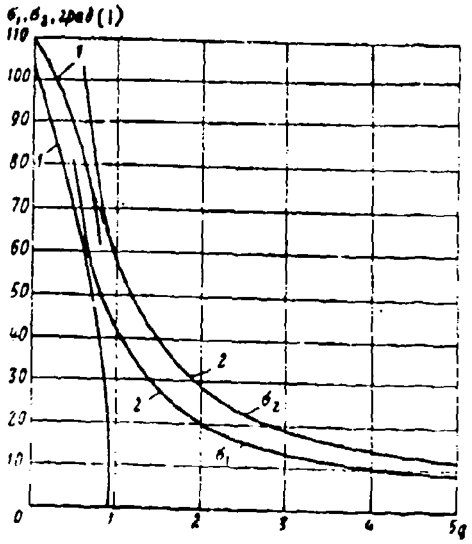

Fig. 2.20. Dependance of the root-mean-square deviation of phase from the relationship/ratio of the voltages of signal and noise: 1 with weak signals; 2 - with powerful signals.

Key: (1) deg.

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With the powerful signals when the distribution of the deviations of phase normal, root-mean-square deviation grow/rises $\sqrt{2}$ once:

$$\sigma_2 = \sqrt{2}\sigma_1 = \frac{1}{q}. \qquad (2.57)$$

Computation of root-mean-square deviation with weak signals gives

$$\sigma_2 = \sqrt{\frac{\pi^2}{3} - \pi q^2}. \qquad (2.58)$$

The dependence of the root-mean-square deviations of phase from the relation of the voltages of signal and noise for the second case is also represented in Fig. 2.20.

A phase difference of signals will cause during measurement offset of arrow/pointer α, but deviations obtained above will cause beat of pointer. As a result of the inertness of measuring meter to it, will operate those spectral components of the deviations which lie/rest within the limits of its band of susceptibility $B_{н}$. Thus, root-mean-square throw of pointer as a result of the effect of noises will be proportional to the root-mean-square deviation of phase and to square root of the relation of passbands after and to phase-meter cell/element [1].

$$\alpha_{ш} = \sigma\sqrt{\frac{2B_{н}}{B_{ф}}}.$$

ENDFOOTNOTE.

[1] This consideration is correct with powerful signals when deviations of phase follow normal law. With weak signals because of a change in the form of the spectrum, the given relationship/ratio is approximated. ENDFOOTNOTE.

Observer will note systematic deviation when

$$a \geqslant a_{ш} C,$$

where C - the coefficient of discernability whose value can be accepted equal to 0.5-1.

Taking into account that

$$q = \frac{E}{E_{пом}} = \frac{E}{\varepsilon_{пом}\sqrt{B_{ш}}},$$

and utilizing (2.55), (2.56), (2.57), (2.58), we obtain the formulas, which determine sensitivity.

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For the first case (one of the oscillations it is free from interferences):

$$E = \left(\frac{\pi\sqrt{\pi}}{6} - \frac{\psi_0^2}{2\sqrt{\pi}C^2}\right)\varepsilon_{пом}\sqrt{2B_{ш}}$$ with weak signals,

$$E = \frac{C\mathcal{E}_{\text{пом}}\sqrt{2B_{\text{н}}}}{\sqrt{2\psi_0}} \quad \text{with powerful signals.} \tag{2.59}$$

For the second case (both oscillations contain interferences):

$$E = \mathcal{E}_{\text{пом}}\sqrt{2B_{\text{н}}}\sqrt{\frac{\pi}{3} - \frac{\psi_0^2}{\pi C^2}} \quad \text{with weak signals,}$$

$$E = \frac{\mathcal{E}_{\text{пом}}\sqrt{2B_{\text{н}}}}{\psi_0}C \quad \text{with powerful signals.} \tag{2.60}$$

The measured phase difference is approximately proportional to the bearing

$$\psi = k_{\text{п}}\theta.$$

Scaling factor $k_{\text{п}}$ can be more than unity. After substituting into formulas (2.59) and (2.60) $k_{\text{п}}\theta$, instead of ψ_0, let us determine the sensitivity of direction finder.

2.13. Sensitivity of two-channel radio direction finder.

The noise voltages in two-channel radio direction finder enter on plate X and Y of cathode-ray tube, as is evident from Fig. 2.13, it is separate from channels 1 and 2 respectively. Noise voltages are the random variables, distributed according to normal law. According to this same law are distributed the deviations of focus along X-axis and Y, proportional to the appropriate voltages.

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In the absence of the signal of the probability of the variation of values x and y, are respectively equal to

$$\left.\begin{aligned} W(x) &= \frac{1}{\sqrt{2\pi}\sigma_x} e^{-\frac{x^2}{2\sigma_x^2}}, \\ W(y) &= \frac{1}{\sqrt{2\pi}\sigma_y} e^{-\frac{y^2}{2\sigma_y^2}}, \end{aligned}\right\} \quad (2.61)$$

where σ_x and σ_y are the root-mean-square deviations, proportional to the appropriate noise voltages.

Probability that the focus will have coordinates x and y, will be defined as combined probability of two events, the probability of each of which is determined by formulas (2.61). It is possible to count that the noise voltages in two channels are statistically independent and combined probability is equal to the product of particular probabilities, i.e.,

$$W(x, y) = W(x) W(y) = \frac{1}{2\pi\sigma_x\sigma_y} e^{-\left(\frac{x^2}{2\sigma_x^2} + \frac{y^2}{2\sigma_y^2}\right)}. \quad (2.62)$$

Let us determine locus, the probability of remaining at which is constant. This will be also the geometric place of the points, mean retention time at which the trace of electron beam is identical. Since the brightness of point on screen is proportional to time of remaining on it of electron beam, locus in question will be curve equal to brightness.

From formula (2.62) it is evident that the probability will be constant, when

$$\frac{x^2}{2\sigma_x^2}+\frac{y^2}{2\sigma_y^2}=K_0^2.$$

This equation of ellipse.

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If channels are completely identical, the effective values of the noise voltages in both channels equal to $\sigma_x=\sigma_y=\sigma$ and ellipse is converted into the circumference

$$x^2+y^2=\rho^2$$

Image on the screen of cathode-ray tube will represent circle with the brightness, maximum in center and which gradually decreases along a radius. If we to plates feed direct/constant voltages, the calling deviations X_0 and Y_0, the center of circle will move into points x_0, y_0. But if we to plates feed the voltages of the signal

$$U_x = U_c\cos\theta\sin\omega t,$$
$$U_y = U_c\sin\theta\sin\omega t,$$

the center of circle will be moved on the straight line whose parametric equations will be

$$x = a\cos\theta\sin\omega t,$$
$$y = a\sin\theta\sin\omega t,$$

and angle of inclination relative to y axis is equal to bearing θ. Image on screen will be approximately rectangle with the luminous intensity, which decreases on perpendicular to its center line. Bearing is counted off on center line of rectangle. Because of indeterminancy/uncertainty (it is more precise, to absence) the boundary of rectangle the reading of bearing is conducted with certain error. The maximum length of rectangle is equal to a. If we the width of rectangle conditionally rate/estimate by value $\rho_ш$, that maximum possible error will be expressed as

$$\Delta\theta_0 \approx \mathrm{tg}\,\Delta\theta_0 = \frac{\rho_ш}{a}.$$

The beam deflections on the screen of cathode-ray tube are proportional to the conducted/supplied to plates voltages, and receiver-amplifier device is linear system. The value of the limit of error can be expressed by the relation of the voltage of signal and noise on the input of the receiving indicator:

$$\Delta\theta_0 = \frac{U_ш}{U_c} = \frac{E_{пом}}{E}. \qquad (2.63)$$

Experiment shows that the mean error composes approximately one the twentieth from angle $2\Delta\theta_0$, if we beyond the arbitrary boundary of the width of strip on screen accept square mean $\rho_ш$.

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2.12. Interference shielding of radio direction finder.

Under the effect on the radio direction finder of two signals, fundamental and mixing, readings of bearing for a fundamental signal can change.

The ability of radio direction finder to preserve within known limits his accuracy in the presence of interference he is called its interference shielding.

The interference shielding of radio direction finder characterizes the noise field intensity which causes the error, which does not exceed the permissible value. The higher the noise field intensity, those, obviously, is better interference shielding. The noise field intensity, which characterizes interference shielding, depends on the strength of the field of signal, value of detuning in jamming frequency relative to signal and from solid angle between the directions of signal and interferences. The noise field intensity usually is assigned relative to the strength of the field of signal (in dB). For the experimental estimate/evaluation of real interference shielding, is required the comprehensive examination/inspection of radio direction finder, in order to establish/install the effect of the indicated factors.

Interference shielding has different character on the levels of the signal and interferences, which do not exceed the linear range of the work of all quasi-linear cell/elements (amplifier tubes and frequency converters), and in cases when the levels of signal and interference exceed the limits of linearity.

Interference shielding in linear (it is more precise, quasi-linear) mode/conditions is determined by resonance receiver response and diagram antenna directivity. Thus far the levels of signal and interference do not exceed limits of linearity, interference shielding barely depends on absolute sound level and signal and depends only on their relative value. The relation of disturbing voltage and signal on output is equal to the same relation at input, multiplied by the value of the relative amplification factor with the assigned detuning, and it is determined by the form of resonance characteristic.

The direction finders at which entire circuit of amplification and transformation of signal can be considered as linear, include auditory and two-channel radio direction finders.

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In auditory radio direction finders with beat reception, the signal and interferences give at output the sound vibrations of difference frequency. The hearing aid of operator distinguishes well the sounds of different tone. The discrimination of two sound signals is facilitated still by the "semantic" selectivity: operator is distinguished them by the character of work, on the sense of the transmitted text, etc. The action of interference does not cause, thus, systematic error. However, powerful interferences mask fundamental signal, they impede the reading of bearing and lead to an increase in the subjective (random) error. With very powerful signals the reading of bearing becomes impossible.

In two-channel radio direction finders with reading on cathode-ray tube, it is represented possible to count off bearings to signal and interferences separately and systematic error because of interference is absent. This phenomenon he is called "visual selectivity" and it is examined in detail in §8.3.

In the radio direction finders of other systems receiving circuit, is substantially nonlinear cell/element - detector.

Let us examine action on the detector of two signals

$$u_1 = A_1 \sin(\omega_1 t + \varphi_1),$$
$$u_2 = A_2 \sin(\omega_2 t + \varphi_2),$$

where ω_1, ω_2, ϕ_1, ϕ_2 are frequencies and the phases of two signals; A_1, A_2 are their amplitudes at the input of detector.

Taking into account the selectivity of receiving circuit, we can record

$$\left.\begin{aligned} A_1 &= U_{1\text{вх}} f(\omega_1), \\ A_2 &= U_{2\text{вх}} f(\omega_2), \end{aligned}\right\} \quad (2.64)$$

where $U_{1\text{вх}}, U_{2\text{вх}}$ are amplitudes of stresses on the input of receiver; $f(\omega)$ - resonance receiver response.

After designating $\omega_2 = \omega_1 + \Delta\omega$, let us find amplitude and the phase of the resulting oscillation:

$$\left.\begin{aligned} u_1 + u_2 &= U \sin(\omega_1 t + \psi), \\ U &= \sqrt{A_1^2 + A_2^2 + 2A_1A_2 \cos(\Delta\omega t + \varphi_2 - \varphi_1)}, \\ \operatorname{tg}\psi &= \frac{A_1 \sin\varphi_1 + A_2 \sin(\Delta\omega t + \varphi_2)}{A_1 \cos\varphi_1 + A_2 \cos(\Delta\omega t + \varphi_2)}. \end{aligned}\right\} \quad (2.65)$$

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Both amplitude and the phase of the resulting voltage vary with beat frequency $\Delta\omega$. If beat frequency lie/rests outside the passband of low-frequency amplifier taking into account the band of reception of indicator, the latter does not reproduce beatings. Readings of indicator, as is evident from (2.65), they depend on amplitude and phase of both component voltages, which leads to the appearance of a systematic bearing error. The course of the computations of error depends on direction-finding method.

Without giving intermediate lining/calculations, we give the results of the computations of the error with different direction-finding methods in the form Table of 2.1. Table is comprised for the square law detector and small relative disturbing voltage ($A_2/A_1 \ll 1$). It should be noted that with small interferences the result of linear detection qualitatively does not differ from the quadratic. [1].

FOOTNOTE [1]. This confirmation is related to linear inertia-free detector. Detector can be considered inertia-free during satisfaction of condition [1.1].

$$\Delta\omega C_{н} R_{н} < \frac{\sqrt{A_1^2 - A_2^2}}{A_2} \approx \frac{A_1}{A_2},$$

where $C_{н}$, $R_{н}$ are a capacitance/capacity and the load impedance of detector. ENDFOOTNOTE.

Besides already stipulated above in the table are accepted the designations:

Δ - an error, is glad; $F(\theta)$ - radiation pattern; $F'(\theta)$ - its derivative in terms of θ; θ_1, θ_2 - bearings of signal and interference; $2\theta_{0.7}$ - width of radiation pattern at the level 0.7; $J_1(x)$ - the Bessel function of first first-order kind; 2b - the distance between two spaced antennas; R - the radius of a circle of antenna location.

From table it is possible to make following conclusions for the radio direction finders, which use the recrified signal:

1. The bearing error, caused by interference, is proportional to the square of the relation of disturbing voltages and signal on the input of detector.

2. Bearing error the lesser, the acute/sharper radiation pattern or more separation of antennas.

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Table 2.1.

(1) Метод пеленгования	(2) Общая формула ошибки	(3) Максимальная ошибка	(4) Угол, при котором ошибка максимальна
(5) По минимуму коэффициента модуляции	$\Delta=\frac{A_2^2}{A_1^2}\frac{F(\theta_2-\theta_1)}{F'(0)}$		
(6) То же при косинусоидальной диаграмме направленности	$\Delta=\frac{A_2^2}{A_1^2}\sin(\theta_2-\theta_1)$	$\Delta=\frac{A_2^2}{A_1^2}$	$\theta_2-\theta_1=90^\circ$
(7) По фазе коэффициента модуляции: а) при косинусоидальной диаграмме направленности	$\Delta=\frac{A_2^2}{A_1^2}\sin(\theta_2-\theta_1)$	$\Delta=\frac{A_2^2}{A_1^2}$	$\theta_2-\theta_1=90^\circ$
б) при острой диаграмме направленности $F(\theta)=\cos^n\theta/2$ с отсчетом по максимуму	$\Delta=\frac{2A_2^2}{A_1^2}\cos^{(2n-1)}\left(\frac{\theta_2-\theta_1}{2}\right)\times$ $\times\sin\left(\frac{\theta_2-\theta_1}{2}\right)$	$\Delta=\frac{A_2^2}{A_1^2}\theta_{0,7}$	$\theta_2-\theta_1=1{,}7\theta_{0,7}$
(8) По фазе высокой частоты (радиопеленгатор с циклическим измерением фазы)	$\Delta=\frac{A_2^2}{A_1^2}\times$ $\times\frac{\cos\left(\frac{\theta_2-\theta_1}{2}\right)J_1\left(2mR\sin\frac{\theta_2-\theta_1}{2}\right)}{mR}$	$\Delta=\frac{A_2^2}{A_1^2}\frac{0{,}58}{mR}$	$\theta_2-\theta_1=\frac{1{,}86}{mR}$
(9) По фазе высокой частоты (интерферометр с неподвижными антеннами)	$\Delta=\frac{A_2^2}{A_1^2}\times$ $\times\frac{\sin\left(\frac{\pi b}{\lambda}\sin\theta_2\right)}{\frac{\pi b}{\lambda}\cos\theta_1\cos\left(\frac{\pi b}{\lambda}\sin\theta_1\right)}$	$\Delta=\frac{A_2^2}{A_1^2}\frac{\lambda}{\pi b}$ (при $\theta_1=0$)	$\theta_2=\frac{\lambda}{2b}$

Key: (1). Direction-finding method. (2). General formula of error. (3). Maximum error. (4). The angle at which the error is maximum. (5). On the minimum of modulation factor. (6). The same with cosinusoidal radiation pattern. (7). On the phase of modulation factor: a) with cosinusoidal radiation pattern; b) with acute/sharp radiation pattern $F(\theta) = \cos^n \theta/2$ with reading of maximum. (8). On the phase of high frequency (radio direction finder with a cyclic measurement of phase of). (9). On the phase of high frequency (interferometer with the motionless antennas).

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3. Direction of arrival of interference, which corresponds to its maximum effect, that nearer to direction of arrival of signal, the acute/sharper radiation pattern or more separation of antennas.

4. Interference shielding depends on method of direction finding. From the examined methods the best interference shielding possesses the method of the cyclic measurement of phase in high frequency.

With powerful interferences begins the overloading of the separate cascade/stages of receiver, which leads to the appearance of combination and crosstalk. In these cases the interference shielding depends on the distribution of selectivity between the separate cell/elements of receiver. This question is examined in the common/general/total courses of radio reception. Some specific phenomena, characteristic to two-channel radio direction finders under the effect of powerful interferences, are examined into §8.3.

The application/use of commutation or modulation of signal at the input of receiver leads to decrease in the interference shielding. Interference just as signal, undergoes commutation or modulation. Spectrum of the switched interference consists of carrier and a series of the side frequencies, distant in frequency from carrier to the values, to the multiple frequency of commutation. The voltage of carrier jamming frequency, if it is sufficiently detuned relative to signal, is attenuate/weakened during the passage of the selective circuits of receiver and at the output of detector can have the permissible low value. At the same time some of the side frequencies can hit the passband of receiver. The voltages of these frequencies they pass through entire circuit of receiver without weakening. Although the voltages of side frequencies on the input of receiver are much lower than the voltage of carrier frequency, at output they can render/show more than weakened by selectivity of the voltage carrying of frequency and exceed the permissible limit. The spectrum of side frequencies the wider, the higher the frequency of commutation, but the intensity by their the greater than less flat is the curve of commutation. With modulation by sine voltage with the preservation/retention/maintaining of carrier the spectrum contains only in one side frequency snizu and on top from carrier. More intense side frequencies appear with commutation with sharp transitions.

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For a decrease in the interferences of commutation, one should apply, when this is possible, commutation with the smooth increase of tension. Is applied also the closing of the input of receiver to time of the course of the transient processes of commutation. In this case, the interferences do not operate on indicator. For the realization of this method, it is necessary that the band of passage of receiver would be considerably (order of 10 times) is more than the frequency of commutation.

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Chapter 3.

ANTENNA SYSTEMS OF RADIO DIRECTION FINDERS.

In radio direction finders are applied the omnidirectional and directional antennas. The fundamental parameters of the antennas which must be determined for the calculation of radio direction finder, are entry impedance, efficiency, to effective height [1]. and directional characteristic.

FOOTNOTE [1]. Besides effective height antenna can be characterized by the effective or absorbing surface ([8.13], h I, page 227-229). ENDFOOTNOTE.

During the determination of bearing, usually is utilized normal-polarized electromagnetic field (vertical electric field of wave). However, for the directional antenna it is necessary to know directional characteristic, efficiency and effective height not only with the reception of normal-polarized electromagnetic field (for the calculation of the effectiveness of radio direction finder), but also the same characteristics with the reception of abnormal-polarized electromagnetic field (for the calculation of the polarizational errors of radio direction finder) (see Chapter 6).

It is sometimes expedient, especially on VHF, to realize direction finding on the horizontal component of electric field.

The simplest antenna systems of radio direction finders have size/dimensions (separation of antennas) less than wavelength. Such systems possess cosinusoidal (figure-of-eight) directional characteristic and are applied in the form of rotary or motionless system.

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Analogously antenna systems with the size/dimensions, greater than wavelength, also are applied as rotatable or motionless systems.

The width of the frequency band of the direction-finding antenna is determined by the facts that at cut-off frequencies they deteriorate, reaching the maximum permissible values, sensitivity and the accuracy of radio direction finder.

Are given below description and the calculation of the parameters of the antennas, used in radio direction finders.

3. 1. The vertical wire antenna.

The simplest antenna is vertical wire. When electromagnetic wave during its propagation reaches antenna, it induces in it emf of high frequency, in consequence of which in antenna begins to circulate the current of the same frequency. This current causes the emission/radiation of the part of the energy back into space. Both processes in antenna - the perception of energy from space and reradiation of this energy - closely interconnected and are, actually, the manifestation of the single ability of antenna to interact with the surrounding field. During the study of the questions of direction finding us interest both processes. On one hand, the vertical wire antennas enter in many instances as component part of the receiving system of radio direction finder, and here is utilized their ability of the perception of energy from the progressive electromagnetic wave. On the other hand, reraliation of antennas (both forming part of radio direction finder and not entering it) produces change in initial electromagnetic field and thereby it can darange of direction finder. The effect of reradiation of the adjacent antennas is examined in chapter 5.

Recall briefly some fundamental properties of the vertical wire antenna (vibrator).

Vibrator can be from the which interests us point of view approximately considered as long line with losses.

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Entry impedance of this line is expressed

$$Z_{вх} = \rho_c \operatorname{cth} \gamma l. \qquad (3.1)$$

where ρ_c is aquivalent wave impedance of the vibrator:

$$\rho_c = \rho_в \left(1 - j\frac{\beta_в}{m}\right);$$

$\rho_в$ - the wave impedance of vertical vibrator; γ - propagation constant:

$$\gamma = \beta_в + jm; \qquad (3.2)$$

$\beta_в$ - decay constant; $m = 2\pi/\lambda$ - is constant of phase displacement.

In practice are applied symmetrical and asymmetric vibrators (Fig. 3.1).

Let us examine the characteristics of symmetrical vibrator. The wave impedance of vertical symmetrical vibrator in free space is designed at small lengths l of vibrator $(l<\lambda)$ by the formula

$$\rho_{\text{в}} = 120\left(\ln\frac{l}{a} - 0{,}69\right). \qquad (3.3)$$

By V. N. Kessenikh is proposed for any l approximation formula [3.1]

$$\rho_{\text{в}} = 120\left(\ln\frac{l}{\pi a} - 0{,}577\right), \qquad (3.4)$$

where a is a radius of vibrator.

If vibrator consists of several (n) wires, arrange/located in circumference with a diameter of D, then the equivalent diameter of vibrator [3.4] will be

$$D_{\text{э}} = D\sqrt[n]{\frac{nd}{D}}, \qquad (3.5)$$

where d is a diameter of wire.

For each ratio D/d, there is a value $n_{\text{макс}}$ greater than which it does not have sense to increase the number of wires of vibrator.

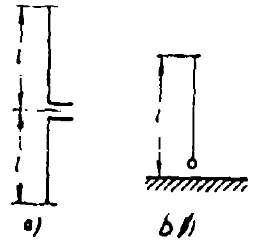

Fig. 3.1. Vibrators: a is symmetrical vibrator; b is asymmetric vibrator.

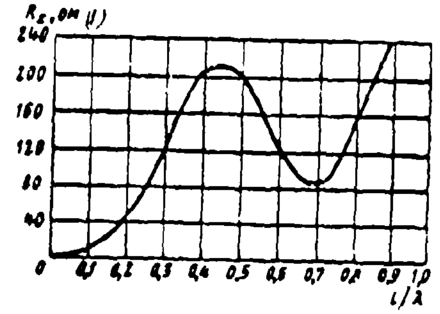

Fig. 3.2. Dependence of radiation resistance of symmetrical vibrator on relation $\frac{l}{\lambda}$.

Key: (1) ohm.

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The capacitance/capacity of vibrator (is linear)

$$C=\frac{1,1}{4\left(\ln\frac{l}{a}-0,69\right)},\ \frac{\text{пф}}{\text{см}}, \qquad (3.6)$$

The attenuation

$$\beta_{\text{в}} = \frac{R_{a}}{\rho_{\text{в}} l \left(1 - \frac{\sin 2ml}{2ml}\right)}, \qquad (3.7)$$

where R_a - the effective resistance, in reference to the current antinode, consists of the radiation resistance and resistor/resistance of the losses:

$$R_a = R_{\Sigma} + R_{\text{пот}} \approx R_{\Sigma}.$$

The dependence R_{Σ} on l/λ is given in the curve of Fig. 3.2.

For low values $l/\lambda < 0.1$, resistor/resistance R_{Σ} is designed from the formula

$$R'_{\Sigma} = 800\left(\frac{h_e}{\lambda}\right)^2.$$

Here R'_{Σ} is referred to current in the middle (at in-feed) of vibrator.

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The effective height of symmetrical vibrator, in reference to current in the loop:

$$h_e = \frac{2\lambda}{\pi} \sin^2 \frac{ml}{2}. \qquad (3.8)$$

The effective height of symmetrical vibrator, in reference to current in its middle:

$$h_e = \frac{2(1 - \cos ml)}{m \sin ml} = \frac{2 \operatorname{tg} \frac{ml}{2}}{m}. \qquad (3.8')$$

When the length of vibrator $2l$ is close to λ, formula (3.8') it is not used, since in such cases it cannot be proceeded from the taken during the derivation of formula (3.8') sinusoidal current distribution along vibrator. With $2l$, close to λ, it is necessary for the calculation effective height to use formula (3.8).

For an asymmetric vibrator wave impedance and radiation resistance will be two times less than at the symmetrical vibrator of the same size/dimensions, i.e., for an asymmetric vibrator it is possible to use the given formulas for R_Σ and $\rho_\text{в}$, by taking into account only coefficient of 1/2.

If we in (3.1) substitute (3.2), then after some transformations we will obtain for entry impedance of the symmetrical vibrator

$$Z_{\text{вх}} = \rho_\text{в} \frac{\operatorname{sh} 2\beta_n l - \frac{\beta_n}{m} \sin 2ml}{\operatorname{ch} 2\beta_n l - \cos 2ml} - j\rho_\text{в} \frac{\frac{\beta_n}{m} \operatorname{sh} 2\beta_n l + \sin 2ml}{\operatorname{ch} 2\beta_n l - \cos 2ml}. \qquad (3.9)$$

More accurate results for entry impedance of vibrator it is possible to obtain, if we in formula (3.9) under signs sin and cos instead of m substitute km, where k depends on $\frac{2l}{\lambda}$ [3.5].

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For $\frac{l}{\lambda}$ = 0-0.35 and 0.65-0.85 formula (3.9) is simplified and assumes the form

$$Z_{вх} = \frac{R_A}{\sin^2 ml} - j\rho_в \operatorname{ctg} ml = R_{вх} + jX_{вх}. \quad (3.10)$$

The value of active and reactive components $Z_{вх}$ for different relations l/λ and different values $\rho_в$ are given in Fig. 3.3.

For the asymmetric vibrator of expression for entry impedance $Z_{вх}$ they are retained, if $\rho_в$ is wave impedance of asymmetric vibrator, l - its length.

When it is required to consider effect on the vibrator of adjacent with it vibrators, they usually use the method of induced emf.

As is known, to account for the mutual effects of vibrators according to the method of those induced by emf to the internal resistance of vibrator, are added the resistor/resistances, introduced adjacent. The resistor/resistance, introduced by any adjacent vibrator in the case when currents in vibrators coincide in amplitude and phase, he is called reciprocal resistance of two vibrators.

Mutual impedance of vibrators is determined from the designed curves X_{12}, R_{12} [3.1, 3.4].

When the length of vibrator is small $(l \ll \lambda)$, it is possible to use the approximation formulas for mutual impedances:

$$\left.\begin{aligned} R_{12} &= \frac{3}{2} R_{\Sigma 0} \left[\frac{\cos md}{m^2 d^2} - \frac{\sin md}{m^3 d^3}(1 - m^2 d^2)\right], \\ X_{12} &= -\frac{3}{2} R_{\Sigma 0} \left[\frac{\sin md}{m^3 d^3} + \frac{\cos md}{m^3 d^3}(1 - m^2 d^2)\right], \end{aligned}\right\} \quad (3.11)$$

where d is a distance between vibrators; $R_{\Sigma 0} = 20 m^2 l^2$ - the radiation resistance of the secluded vibrator by length l; R_{12} and X_{12} are referred to current of foundation.

During the calculation of coupled impedances, one should consider the possible dissimilarity of values and phases of currents in separate vibrators.

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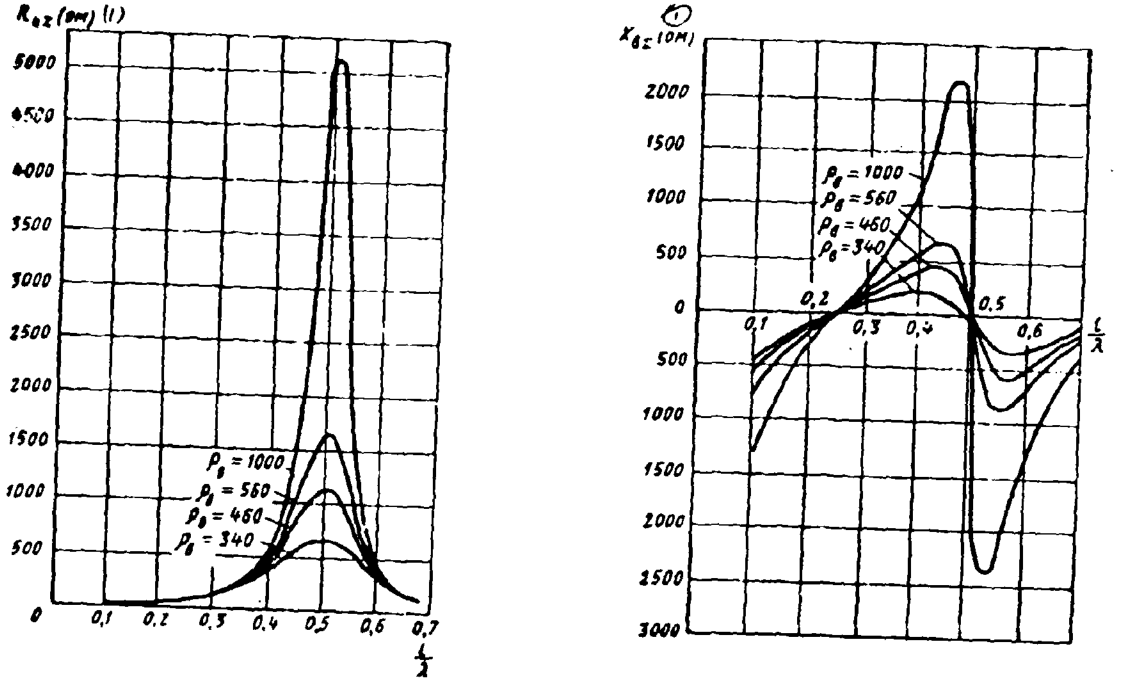

Fig. 3.3. Dependence $R_{вх}$ and $X_{вх}$ symmetrical vibrator on relation $\frac{l}{\lambda}$ with different $\rho_в$.

Key: (1) (ohm).

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Impedance, introduced into any (n-s) antenna, in the general case will be:

$$\begin{gathered}k_1 R_{n1}\cos\psi_1+k_2 R_{n2}\cos\psi_2+k_3 R_{n3}\cos\psi_3+\ldots\\ \ldots-k_1 X_{n1}\sin\psi_1-k_2 X_{n2}\sin\psi_2-k_3 X_{n3}\sin\psi_3-\ldots\\ \ldots+j\,(k_1 R_{n1}\sin\psi_1+k_2 R_{n2}\sin\psi_2+k_3 R_{n3}\sin\psi_3+\ldots\\ \ldots+k_1 X_{n1}\cos\psi_1+k_2 X_{n2}\cos\psi_2+k_3 X_{n3}\cos\psi_3+\ldots)=\\ =R_{\text{внес}}+jX_{\text{внес}}.\end{gathered}$$

where R_{n1}, R_{n2}, R_{n3} - the active components of mutual impedances of the n antenna with the 1st, 2nd, 3rd so forth antennas; X_{n1}, X_{n2}, X_{n3} - reactive components of the same resistor/resistances; k_1, k_2, k_3 - an amplitude ratio of the currents of the n antenna and the 1st, 2nd, 3rd of antennas; ψ_1, ψ_2, ψ_3 - the lead angles of the phase of current into the 1st the 2nd and the 3rd antennas with respect to the phase of current in the n antenna.

In horizontal plane the vertical wire antenna is not directed, i.e., by the antenna of circular action.

Expression for the radiation pattern of symmetrical vibrator in vertical plane will be [3.4]

$$\begin{gathered}E_c=\frac{2E}{m}\frac{\cos(ml\sin\beta)-\cos ml}{\cos\beta}\times\\ \times\sqrt{1+R_1^2+2R_1\cos(\varphi_1-2mH\sin\beta)}.\qquad(3.12)\end{gathered}$$

For an asymmetric vibrator radiation pattern in vertical plane is expressed

$$\begin{gathered}E_{nc}=\frac{E}{m\cos\beta}\,\{[\cos(ml\sin\beta)-\cos ml]\,(1+R_1\cos\varphi_1)+\\ +R_1\sin\varphi_1\,[\sin(ml\sin\beta)-\sin ml\sin\beta]\}+\\ +j\,\{[\sin(ml\sin\beta)-\sin ml\sin\beta]\,(1-R_1\cos\varphi_1)+\\ +R_1\sin\varphi_1\,[\cos(ml\sin\beta)-\cos ml]\},\qquad(3.13)\end{gathered}$$

where β is an angle of the slope of a front of wave; R_1 - the module/modulus of the coefficient of terrain echo; Φ_1 - the argument of the same coefficient; H is a height/altitude of the center of the vibrator above the earth/ground; E - the strength of field in free space.

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The form of the radiation patterns of the grounded vertical vibrator of different length with the ideal conducting earth/ground is given in Fig. 3.4.

The range of the use of a vibrator is determined by the frequency properties of the radiation pattern and entry impedance.

With the shortening of wavelength, i.e., an increase of the relation l/λ, in vertical radiation pattern appear minor lobes (Fig. 3.4). For this reason relation $\frac{l}{\lambda}$ they limit by value $\frac{l}{\lambda_{мин}}$ = (0.5-0.625).

From Fig. 3.3 it follows that entry impedance of vibrator changes in the fact limit inferiors, the lesser the wave impedance of vibrator. Wave impedance decreases with an increase in the radius of cross section, i.e., during the use of thick vibrators. As can be seen from (3.5), the same result can be obtained if vibrator consists of several wires of a small diameter, arrange/located in large-diameter circumference. The essential expansion of the band coverage of vibrator is obtained also during the use of vibrators of conical or exponential shape (Fig. 3.5).

The wave impedance of the conical vibrator

$$\rho_{\text{к}} = 138 \lg \left(\operatorname{ctg} \frac{\psi}{2} \right),$$

where 2ψ is a cone apex angle.

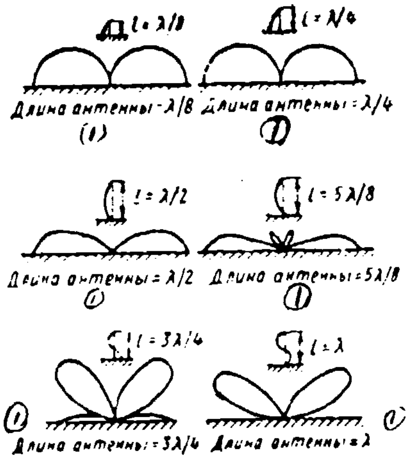


Fig. 3.4. Vertical directional characteristic of the vertical grounded vibrator depending on its length.

Key: (1). Length of antenna.

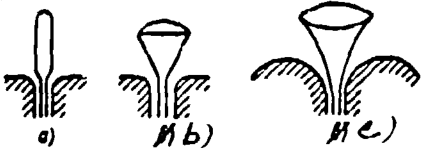
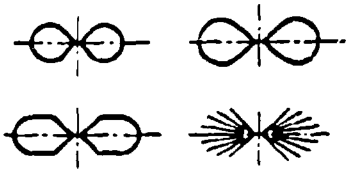


Fig. 3.5. Broadband of vibrator (asymmetric): a - cylindrical; b is conical; c is exponential.

Fig. 3.6. Broadband planar symmetrical vibrators.



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In symmetrical fulfillment antenna he is called biconical.

The calculation of entry impedance of conical antenna here is not given. For the connection of antenna, it is expedient to use feeder with wave impedance $\rho_\Phi = \rho_K$. Investigations show ([3.3], page 107) that for $\psi \geqslant 30^\circ$ KBV > 0.5 during satisfaction of condition $\frac{2\pi l}{\lambda_{\text{макс}}} \leqslant$ 1.2, where l is length forming of cone, i.e., $\lambda_{\text{макс}} \approx 5l$. Minimum transmitting wave is limited to appearance in vertical directional characteristic of the deep minimums feast the low values of angle β. So that this would not be, it is required to fulfill $\lambda_{\text{мин}} \approx 0{,}8l$. Thus, the frequency band, overlapped by conical antenna, $\geqslant 5 \div 6$.

One Of the cones of the biconical antenna can be replaced by disk. Discone antenna has overall sizes less than biconical; frequency band of its is somewhat less.

The value of wave impedance depends on the maximum size of cross section, which makes it possible to utilize planar constructions of vibrators (Fig. 3.6), that have rectangular cross section. Planar vibrator can be made also from separate conductors.

Page 86. Application/use wide-range of the vibrators of the indicated types frequently causes difficulties due to their large overall sizes. The range of the use of a fine/thin vibrator can be expanded by the inclusion into it of reactive and active cell/elements. An example of this antenna is given in Fig. 3.7. The cell/elements of antenna are selected from following considerations [7.12].

Bierat $l_1<\frac{\lambda_{мин}}{4}$ and $l_2<\frac{\lambda_{мин}}{4}$ with those, so that these cuttings off would be capacitance. Then l_1 and l_2 it is possible to replace with the equivalent capacitors C_1 and C_2 respectively.

For that frequency $\omega_1/2\pi$ by which chain/network L_1C_1 is inclined into resonance ($\omega^2{}_1L_1C_1 = 1$), entry impedance at point 2 will be equal to the wave impedance ρ_3 of section l_3, if $R_1 = R_2 = \rho_3$ and $L_2 = \rho_3^2 C_2$. In this case, on shortest wave, antenna radiation is determined virtually only by section l_3. Taking into account the requirement for the absence of minor lobes in radiation pattern, one should accept $l_3 \approx \frac{\lambda_{мин}}{2}$. On longest wave in emission/radiation, participate the sections l_2 and l_3. Is expedient to select $l_2+l_3=\frac{\lambda_{макс}}{4}$, in order to ensure not too low a radiation resistance.

During satisfaction of these conditions

$$\lambda_{макс}=4\,(l_2+l_3)\approx\lambda_{мин}+2\lambda_{мин}=3\lambda_{мин},$$

i.e. the antenna provides the overlap of triple wave band with satisfactory indices.

Recently as range vibrator is widely applied shunt vibrator. Satisfactory agreement this vibrator provides within limits from $\frac{l}{\lambda_{\text{макс}}} \approx$ $\approx 0.16-0.17$ to $\frac{l}{\lambda_{\text{мин}}} \approx 2$. Considerations according to calculation see [3.4].

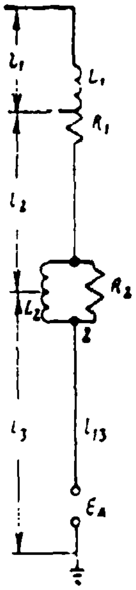

Fig. 3.7. Wide-range antenna with the connected cell/elements L, R.

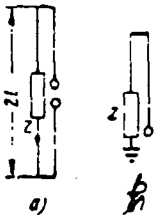

Fig. 3.8. Loop antenna: a is symmetrical; b is asymmetric.

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In chapter 7, are described the methods of the expansion of the working frequency band of the vibrator by applying the cell/elements of agreement or compensation for reactance.

3.2. Loop-Antenna.

Loop-antenna is schematically represented in Fig. 3.8. Are applied symmetrical and asymmetric loop-antennas. Symmetrical loop-antenna consists of two parallel, connected on end/leads conductors, arrange/located at small (less than 1/10-1/20 wavelengths) distance from each other. Are distinguished aperiodic and resonance loop-antennas. In aperiodic loop-antenna into middle of one of the conductors, is included resistor/resistance Z, while in middle of the other - the input of receiver or feeder. Asymmetric loop-antenna is the half of symmetrical. The second half is supplemented by the image of the first in the earth/ground.

If resistor/resistance Z - active is equal to wave loop resistance фа $\rho_{шл}$, that entry impedance equal to wave impedance $\rho_{шл}$ over a wide range of waves (when $\lambda > 4l$). This property of antenna is its key advantage, since constancy and the active character of its entry impedance make it possible to ensure good agreement of antenna with feeder over a wide range of frequencies.

Deficiency/lacks in loop-antenna are small efficiency and decrease of effective height during an increase in the wavelength.

In resonance loop-antenna $\left(\frac{l}{\lambda} = 0.25\right)$ resistor/resistance Z is equal to zero. Its radiation pattern the same as in simple vibrator, radiation resistance into four, but effective height into two grooves is more than the corresponding values of simple vibrator.

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With Z = 0, loop-antenna of the heavy-gauge wires or of the tapes possesses considerably larger band coverage (i.e. by the section of the frequencies where X_a is close to zero and R_u is little affected), than usual dipole.

In radio direction finders loop-antenna can be used as cell/element by the antenna of system (for example, in the spaced antennas). The greatest application/use of a loop-antenna was obtained in ultrashort-wave range.

1.3. Loop antennas.

One Of the widespread types of antenna, used for radio traffic, is the loop antenna (framework), in the simplist form which is the fine/thin conductor, that has the form of the locked plane figure.

Electromotive force within a framework of a small size/dimension.

Let us examine the framework of a small size/dimension, i.e., let us assume that the perimeter of the framework is very small in comparison with wavelength and that over entire length of the framework the current has constant amplitude.

This framework is equivalent to the magnetic dipole, directed along the normal to the framework. If the framework with an area of S is located in electromagnetic field with magnetic field strength H, then the magnetic flux, which penetrates the framework, will be

$$\Phi = (\vec{H}\vec{n}_1)\mu S = H\mu S\cos\varphi,$$

where n_1 is the unit vector of standard to the framework; ϕ - the angle between unit vector and magnetic intensity H, μ - magnetic permeability within the framework.

Emf, induced within the framework, will be

$$E = -\frac{d\Phi}{dt} = -\left(\frac{d\vec{H}}{dt}\vec{n}_1\right)\mu S. \qquad (3.14)$$

Emf that induce in the electric dipole (Hertz doublet) by length l by electrical field E, is equal to

$$\dot{E} = (\vec{E}\vec{l}).$$

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From the comparison of these formulas, it is evident that the dependence of emf of the framework on the three-dimensional/space location of its axis of relatively magnetic field the same as dependence of emf of the electric dipole on its location of relatively electric field. This analogy escape/ensues of the equivalency noted above of the framework to magnetic dipole.

Under the harmonic law of a change in the magnetic field, the operation of differentiation is equivalent to multiplication by $j\omega$ and formula (3.14) passes in

$$\dot{E} = -j\omega S\mu(\vec{H}\vec{n}_1) = -j\omega S\mu H\cos\varphi. \qquad (3.15)$$

If the framework is located in the remote zone (zone of emission/radiation) of transmitter, magnetic intensity can be expressed through the electric intensity

$$H = \frac{E}{120\pi}. \qquad (3.16)$$

After substituting (3.16) in (3.15) and after replacing in it ω by $(2\pi 3\cdot 10^8/\lambda)$ and μ (for air) by $4\pi 10^{-7}$, we will obtain

$$\dot{E} = -j\mathrm{E}\frac{2\pi S}{\lambda}\cos\varphi. \qquad (3.17)$$

From this formula it is evident that emf within the framework lags on phase for 1/4 periods behind the strength of field. The maximum value E occurs with $\phi = 0$ and it is equal

$$E_{\text{макс}} = E\frac{2\pi S}{\lambda}. \qquad (3.18)$$

The effective height of the framework

$$h_e = \frac{E_{\text{макс}}}{E} = \frac{2\pi S}{\lambda}. \qquad (3.19)$$

The law of a change in emf during the rotation of the framework is determined depending on the angle between standard to the plane of the framework and direction of the magnetic field. Virtually in the work of the framework as direction-finding antenna to conveniently have the explicit dependence of emf on the direction of incident wave (sense of the vector of Poynting) with the determined location of the framework.

~~End section~~.

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Let us examine the framework whose plane it is vertical, and axis forms angle θ with X-axis of rectangular coordinate system. In plane ZX, is propagated plane electromagnetic wave, the magnetic intensity of which is determined by formula (2.3); in it it is accepted with r = 0

$$\vec{H} = (-\vec{n}H_\perp e^{j\Phi_n} + \vec{a}H_n)e^{j\Phi}.$$

Vectors n and a have components:

$$n_x = \sin\beta;\ n_y = 0;\ n_z = \cos\beta;\ a_x = 0;\ a_y = 1;\ a_z = 0.$$

We will use formula (3.15).

Vector n_1 has components:

$$n_{1x} = \cos\theta;\ n_{1y} = \sin\theta;\ n_{1z} = 0.$$

The polarized in plane incidence/drops and perpendicular to it components $H_n \overset{\text{and}}{a} H_\perp$ are equal to

$$H_\perp = H\sin\gamma;\quad H_n = H\cos\gamma.$$

where γ is an angle of rotation of the plane of polarization, equal to the angle between vectors of electric field and the vertical plane, which contains direction of propagation.

Fulfilling the scalar multiplication H and n_s in formula (3.15), we obtain

$$\dot{E} = -j\omega\mu SH(-\sin\beta\sin\gamma\, e^{j\psi_e}\cos\theta + \cos\gamma\sin\theta)\, e^{j\psi}.$$

Reject/throwing unessential for future reference phase factor and again passing to electric field, we obtain

$$\dot{E} = -j\frac{2\pi S}{\lambda}E(\sin\theta\cos\gamma - \sin\beta\sin\gamma\cos\theta e^{j\psi_e}). \qquad (3.20)$$

If field either is normally polarized ($\gamma = 0$) at any angle of incidence, or it is propagated horizontally ($\beta = 0$) during any polarization, electromotive force within the framework it is proportional to sin θ and radiation pattern has a form of eight [1] (Fig. 2.5). Let $\beta = 0$, then

$$\dot{E} = -j\frac{2\pi S}{\lambda}E\cos\gamma\sin\theta.$$

FOOTNOTE [1]. If angle θ is is counted off between the plane of the framework and plane ZX of the propagation of electromagnetic wave, then the radiation pattern of the framework is proportional to cos θ.

Therefore it was called the name cosinusoidal directional characteristic. ENDFOOTNOTE.

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By rotation of the framework we can attain zero emf with $\theta = 0$. In this position the standard to the plane of the framework indicates the true direction of wave.

With it is abnormal, but to the linearly polarized waves $(\psi_a = 0)$, which falls with certain slope/inclination, we will obtain

$$\dot{E} = -j\frac{2\pi S}{\lambda} E \sqrt{\sin^2\gamma \sin^2\beta + \cos^2\gamma} \sin(\theta - \Delta). \quad (3.21)$$

where

$$\operatorname{tg}\Delta = \operatorname{tg}\gamma \sin\beta. \quad (3.22)$$

Radiation pattern retains the form of eight, but the direction of zero reception/procedure is obtained when $\theta = \Delta$, i.e., it indicates the direction of the arrival of wave with the error, determined by formula (3.22).

In the case of circular polarization $\left(\psi_a = \frac{\pi}{2}, \gamma = \frac{\pi}{4}\right)$ from (3.20)

$$|E| = \frac{\sqrt{2}\pi S}{\lambda} E \sqrt{\sin^2\theta + \cos^2\theta \sin^2\beta}.$$

Emf to the framework non-vanishing in which position of the framework. With $\theta = 0$ we obtain the minimum of emf

$$E_{\text{мин}} = E \frac{\sqrt{2}\pi S}{\lambda} \sin\beta,$$

while at $\theta = 90°$ - the maximum

$$E_{\text{макс}} = E \frac{\sqrt{2}\pi S}{\lambda}.$$

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With the reading of bearing for audition, is observed the diffuse minimum (angle of the equal to audibility), but bearing error-free.

In the general case of elliptical polarization, is observed both error and the diffuseness of the minimum.

Electromotive force within a framework, which consists of several turns.

The electromotive forces, induced within the framework, are very small. it is expedient to connect several such turns consecutively. If in this case they desire so that all turns would remain in one plane, then the framework acquires the form of spiral (Fig. 3.9).

Let us suppose that the reverse/inverse effect of each turn on the incoming field is so small that it can be disregarded, and the overall length of wire considerably less than the wavelength. Then we can each turn examine independently, also, to each apply formula (3.17); set/assuming $\phi = \pi/2-\theta$, $\beta = 0$, we have

$$\dot{E}_1 = -jE\frac{2\pi S_1}{\lambda}\sin\theta,\quad \dot{E}_2 = -jE\frac{2\pi S_2}{\lambda}\sin\theta \text{ etc.}$$

Here E_1, E_2 ... - emf in each turn.

Resulting emf will be equal to the sum of emf E_1, E_2 and of so forth, induced of the separate turns:

$$\dot{E} = -j\frac{2\pi}{\lambda}E\sin\theta\sum_{k=1}^{N}S_k = -j\frac{2\pi S_{cp}N}{\lambda}E\sin\theta,$$

where S_k - an area of the k turn; $S_{cp}=\frac{\sum_{k=1}^{N}S_k}{N}$ - a middle area of turn; N is a turn number.

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The effective height of the framework

$$h_e = \frac{2\pi S_{cp}N}{\lambda} \qquad (3.23)$$

will be in N of times more than the effective height of one mean turn. The framework can have not only spiral form. If it consists of several turns, it it is possible to wind over the lateral surface of cylinder. In this case all turns are obtained identical size/dimension, but are arrange/located they no longer in one plane, and therefore this framework is called three-dimensional/space (Fig. 3.10).

Under those assumptions that they were made for the spiral framework, we here will obtain for the sum of emf, induced of the turns of the framework, the formula

$$\dot{E} = -j\frac{2\pi NS}{\lambda}E\sin\theta. \qquad (3.23')$$

In the case of the three-dimensional/space framework, one should take into consideration that the wire of the winding of the framework forms one complete turn in the plane, perpendicular to the plane of the fundamental turns of the framework. more graphically anything this can be seen, if we design the framework on this perpendicular to its turns plane [i.e. to look to framework from the side (Fig. 3.11)]. The area of turn, perpendicular to fundamental turns, in the figure is shaded.

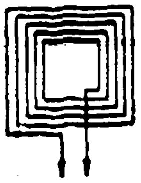

Fig. 3.9

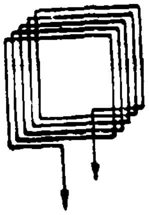

Fig 3.10

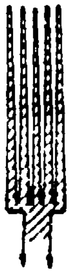

Fig 3.11

Fig. 3.9. Multiturn spiral framework.

Fig. 3.10. Three-dimensional/space framework.

Fig. 3.11. Form of framework on the side.

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Calling this area $S_{\text{н}}$ we obtain emf, induced in the perpendicular turn:

$$\dot{E}_{\text{н}} = -j\frac{2\pi S_{\text{н}}}{\lambda}\mathrm{E}\cos\theta,$$

and total emf within the framework will be

$$\dot{E} = -j\mathrm{E}\frac{2\pi}{\lambda}(NS\sin\theta + S_{\text{н}}\cos\theta) =$$

$$= -j\mathrm{E}\frac{2\pi}{\lambda}\sqrt{(NS)^2 + S_{\text{н}}^2}\sin(\theta + \Delta),$$

where

$$\operatorname{tg}\Delta = \frac{S_n}{NS}. \qquad (3.24)$$

The effective height of this framework

$$h_n = \frac{2\pi}{\lambda}\sqrt{(NS)^2 + S_n^2}.$$

Since S_n usually is considerably less than NS, S_n it is possible to disregard and for the three-dimensional/space framework to apply the formula

$$h_e = \frac{2\pi NS}{\lambda}.$$

It is substantial to note that emf is turned in this case into zero not with $\theta = 0$ as for the flat/plane framework, but when $\theta = \Delta$, i.e., at angle, on Δ arc tg $\frac{S_n}{NS}$ different from the true. On the elimination of the lateral reception/procedure of the framework, see § 4.4.

Effect of nonuniform current distribution.

In radio direction finders are applied the framework of a small size/dimension; however, sometimes size/dimensions are not so small so that to it would be possible completely disregard the nonuniformity of current along the wire of the framework.

Considering that the inductance and the capacitance/capacity of the framework are distributed evenly on it last, i.e., that inductance L_1 and capacitor C_1 per the unit of length are constant, the framework (Fig. 3.12a) can be likened to long line (Fig. 3.12b), also, for research on the process of electromagnetic vibrations in it to use the conclusions of the theory of long lines.

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Distribution along the wire of current I_x and of a potential difference U_x between the appropriate points on two halves of the framework is subordinated to the equations:

$$\left.\begin{aligned} I_x &= I_0 \cos\frac{2\pi x}{\lambda}, \\ U_x &= U_0 \sin\frac{2\pi x}{\lambda}, \end{aligned}\right\} \qquad (3.25)$$

a the wavelengths, which correspond to the natural resonance of the framework, are determined by the expression

$$\lambda_0 = \frac{4l}{k},$$

where k is the whole number.

Greatest of these waves

$$\lambda_0 = 4l \qquad (3.26)$$

is called the natural wavelength of loop antenna. The virtually natural wavelength of the framework is greater than it is obtained according to equation (3.26), as a result of the nonuniformity of the distribution of capacitance/capacity and inductance along the length of the framework.

The current distribution and potential also not completely accurately does follow equations (3.25). However, in the first approximation, it is possible to use these equations.

Impedance on terminals a-b of the single-turn framework is equal

$$Z = j\rho \operatorname{tg} \frac{2\pi l}{\lambda}, \qquad (3.27)$$

where $\rho = k_0 \sqrt{\frac{L_1}{C_1}}$ is wave impedance of the framework; inductance, H•cm^{-1}; capacitance/capacity $\text{ф}\cdot\text{см}^{-1}$; k_0 - the coefficient, depending on the type of the winding of the framework.

During great lengthening, i.e., when $\lambda > \lambda_0$, in expression (3.27) tangent can be replaced it with the argument

$$Z = j\rho \frac{2\pi l}{\lambda} = j\omega L_1 l = j\omega L. \qquad (3.28)$$

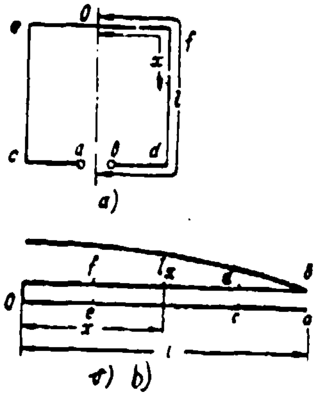

Fig. 3.12. Framework and equivalent to it line.

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Consequently, during great lengthening the framework can be considered as lumped inductance whose value is equal to its static value. During a decrease in the wavelength the framework one should represent in the form of parallel connection of inductance and capacitance/capacity. The values of inductance and capacitance/capacity in duct with the concentrated constants, equivalent to the framework, they are called dynamic inductance ($L_{д}$) and dynamic capacitance/capacity ($C_{д}$). On the strength of formula (3.28)

$$L_{д} = L.$$

From comparison (3.26) with formula for the natural wavelength of duct with the concentrated constants

$$\lambda_{\text{л}} = 2\pi\sqrt{L_{\text{д}}C_{\text{д}}}\,3\cdot 10^{10} = 4l$$

we find for the single-turn framework

$$C_{\text{л}} = \frac{400}{9\cdot\pi^2}\,\frac{l^2\,\text{м}^2}{L\,\text{мкгн}},\ \text{пф}. \qquad (3.29)$$

The radiation resistance of the framework can be found as radiation resistance of Hertz doublet with the effective height, determined by formula (3.23).

Resistance of the emission/radiation of the framework is obtained negligibly small, and it can be usually disregarded in comparison with the resistor/resistance of losses. Formulas for the calculation of the parameters of the framework are given in appendix I.

To account for the effect of nonuniform current distribution on the work of the framework, let us examine the rectangular framework [1] with sides a and b.

FOOTNOTE [1]. During nonuniform current distribution in the framework of emf, induced in it, it depends on the form of the framework. This difference is small, and the obtained further conclusions approximately can be propagated to the framework of any form. ENDFOOTNOTE.

Figures 3.12b shows current distribution in the framework in accordance with equation (3.25).

In the vertical sides of the framework, the current distribution virtually is evenly and, that especially important, is equal in the right fd and left ec sides.

Let us estimate the effect of the dissimilarity of currents in the upper and lower sides of the framework. For this, is increased the current at each point $(l-x)$ of lower side, so that it would become equal to the current in the symmetrical point (x) of upper side.

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It is natural that subsequently we must consider the emission/radiation of current, differing and opposite by sign added. The framework with the "adjusted" current on its emf is equivalent previously examined framework with uniform current distribution, and to it are used all already obtained conclusions. The supplementary current, which takes place in one of the horizontal sides of the framework, can be likened to Hertz doublet. We can consider therefore that the action of vertical sides is retained the same as during uniform current distribution.

In upper and lower sides the currents are essentially different. Thus, the framework with nonuniform current distribution is equivalent to the same framework with uniform current distribution and to the horizontal Hertz doublet. Horizontal Hertz doublet will participate in the reception/procedure only of horizontal component of electric field, i.e., by abnormally polarized field component.

Its effective height is very small and usually the less effective height of the framework. For example, for the square framework (h = b) the ratio of the effective height of dipole $h_{\text{д}}$ to the effective height of the framework $h_{\text{р}}$, is approximately equal

$$\frac{h_{\text{д}}}{h_{\text{р}}} = \frac{3\pi b}{\lambda} \approx 1,2\frac{\lambda_0}{\lambda}$$

and, in the usual relations of λ_0/λ = 0.15-0.5, is 0.18-0.60.

The action of supplementary dipoles especially substantially in the special circuit diagrams of the framework, intended for the exception/elimination of the reception/procedure of the horizontal component of field (see § 37). Remanent/residual harmful reception/procedure will be determined precisely by their action.

3.4. Shielded framework.

Screening of framework consists of the metal tube, bent in the form of circle (or square) within which is placed the winding/coil. For the preservation/retention/maintaining of the possibility of reception/procedure a tube- screen it must be sectional across (Fig. 3.13). Otherwise occurs complete shadowing and reception/procedure it is absent.

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Such construction of the framework provides first of all high mechanical its quality - the strength, watertightness, the protection of winding, especially important in marine and aviation practice. Furthermore, the application/use of a screen provides the symmetry of the winding of the framework and contributes to the exception/elimination of antenna effect (see § 4.2).

Under the influence on the screen of the framework of plane electromagnetic wave in it, is induced (as within the usual framework) the electromotive force, equal to

$$E_3 = E\frac{2\pi S}{\lambda},$$

where S - an area of the bent tube of screen (counting on its centerline), is equal to the middle area of the turn of winding: λ is a wavelength.

If screen is closed, in it appear the currents whose action completely compensates for within screen applied field, and reception/procedure to the framework is absent. the presence of cut/section in screen excludes the possibility of the course of the ring currents of conductivity in screen. Thus, on gap is created potential difference, virtually equal emf of screen E_3, since bias current through the gap is very low and a voltage drop across the inductance of screen can be disregarded, as is evident from the equivalent circuit diagram of screen (Fig. 3.14). the strength of field in gap will be

$$E_3 = \frac{E_3}{d_3},$$

where d_3 is a gap length.

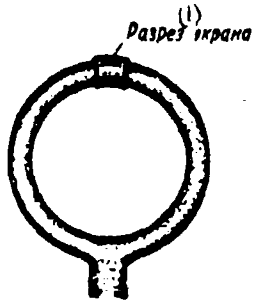

Fig. 3.13

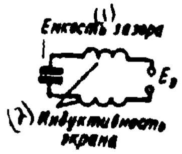

Fig. 3.14

Fig. 3.13. Shielded framework.

Key: (1). Cut/section of screen.

Fig. 3.14. Equivalent diagram of screen.

Key: (1). Gap capacitance. (2). Inductance of screen.

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In each conductor, that is located in gap, it will be induced by emf

$$E_1 = E_{з}d_{з} = E_{з}.$$

Total emf within the framework, if its turn number is equal to N, will be

$$E = NE_1 = \frac{2\pi SN}{\lambda}E.$$

Consequently, the effective height of the shielded framework is equal to its effective height without screen.

Without change remains the inductance of the framework. The shielded framework is the system of two connected lines: the wire of the framework and the internal surface of screen, the external surface of screen - the earth/ground. Because of the communication/connection of lines, the natural wavelength of system is greater than the natural wavelength of the wire of the framework. grow/rises, therefore, the equivalent capacitance/capacity of the framework, and also the effective resistance of the framework because of losses in screen.

Conclusions indicated above are accurate, while the natural wavelength of screen is considerably less than the working wavelength. Thus, for instance, with artificial increase of gap capacitance voltage on it will differ significantly from emf, which operates in the circuit of screen, and it will be considerably more in the case of the tuning of the circuit of screen into resonance. in this case, the value of the inductance of the framework grow/rises. in practice the tuning of gap does not find application/use in view of the fact that usually the work is maintained not on the fixed/recorded wave, but in certain frequency band. Structural/design execution of screen can differ from that which was described above.

3.5. Framework with ferromagnetic core.

For an increase in the effective height of the framework within it can be placed the core from material with magnetic permeability, which exceeds unit.

Let us examine the core, having the form of ellipsoid of revolution and placed into uniform (constant value) applied field in such a way that the major axis of ellipsoid coincides with the direction of magnetic intensity (Fig. 3.12).

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The magnetic induction B and the strength of field H in core have within ellipsoid constant values and one and the same direction, which coincides with the direction of the strength of applied field. The value of magnetic induction is determined by the formula

$$B = \frac{\mu}{1 + (\mu - 1)\varkappa} H_0 = \mu_{д} H_0,$$

where μ is magnetic permeability of core;

H_0 - intensity of applied field (when from the form of body; $\varkappa$ is the demagnetization coefficient, depending on the form of body; $\mu_\text{л}$ is the operating magnetic permeability.

For an ellipsoid of revolution

$$\varkappa = \frac{1-\varepsilon^2}{\varepsilon^3}\left(\frac{1}{2}\ln\frac{1+\varepsilon}{1-\varepsilon}-\varepsilon\right),$$

where $\varepsilon = \sqrt{\frac{b^2-c^2}{b^2}}$ is eccentricity of ellipsoid; b and c - the large and semiminor axes of ellipsoid.

With small eccentricities

$$\varkappa \approx \frac{1}{3}-\frac{2}{15}\varepsilon^2+\frac{2}{35}\varepsilon^4+\ldots$$

For the sphere

$$\varepsilon = 0 \text{ and } \varkappa = \frac{1}{3}.$$

For the strongly elongated ellipsoids

$$\varkappa \approx 2(1-\varepsilon)\left(\frac{1}{2}\ln\frac{1}{1-\varepsilon}-1\right)=\frac{1}{k^2}(\ln 2k-1),$$

where k = b/c.

To an increase in magnetic permeability μ_a once corresponds an increase in the flow of magnetic induction how many times.

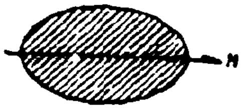

Fig. 3.15. Ellipsoid in uniform magnetic field.

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If we do not consider the phenomenon of eddy currents in core, it is possible to count that the same increase of the flow will be observed in the case of variable field H_0. Electromotive force in the winding, placed on core, also grow/rises $\mu_д$ once. It is possible to count that the effective height of the framework with ferromagnetic core grow/rises $\mu_д$ once. Is introduced also the concept of the effective diameter of the framework, i.e., the diameter of that air framework whose effective height is equal to the effective height of this framework:

$$D_{\text{рд}} = \sqrt{\mu_{\text{д}}}\, D_{\text{р}}.$$

During the practical fulfillment of the ferromagnetic framework to core shape in the form of cylinder with circular or elliptical cross section. In the first approximation, cylindrical core by length l and by cross section S can be for calculations replaced ellipsoid on the condition that $l = 2b$ and $S = \pi c^2$.

The inductance of the framework with ferromagnetic core also grow/rises approximately μ_a once.

If we assign the space (or weight) of core, then with increase in $k = b/c$ its cross section will decrease. On the other hand, μ_a will grow/rise, since with increase in k the demagnetization coefficient decreases. The combined action of these two factors is led to the fact that there is the optimum value k whose value depends on the permeability of the material of core and grow/rises with an increase μ.

With the used at present materials the optimum sense $k = b/c$ lie/rests within the limits:

$k_{\text{опт}} = 3 \div 5$ for the magnetic dielectrics

and

$k_{\text{опт}} = 15 \div 25$ for ferrites.

During the application/use as material of a core of usual electrical sheet steel for the high frequencies at which work the radio direction finders, appear very large eddy currents. The parameters of the framework sharply deteriorate, in particular, noticeably it descends its quality.

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For this reason as material for the cores of the framework, are applied the pressed powder-like ferromagnetic materials and the ferrites.

The major advantage of the framework with ferromagnetic core are its small size/dimensions in comparison with the size/dimensions of the air framework.

3.6. Reception/procedure to two spaced antennas.

One Of the fundamental tops of direction-finding antennas is system of two spaced antennas with the use of differential emf. Radiation pattern of its is determined by the formula, analogous (2.13):

$$\dot{E} = -j2Eh_d F_1(\theta, \beta)\sin\left(\frac{2\pi b}{\lambda}\sin\theta\cos\beta\right), \qquad (3.30)$$

where h_d - the effective height of each of the spaced antennas; $F_1(\theta, \beta)$ - directional characteristic of each antenna; 2b - the distance between antennas; θ is an angle between standard to the line of the connecting antenna and direction of propagation.

This formula considers the action of vertical component of electric intensity. Under the effect of the horizontal component of field, the effective height and directional characteristic of each antenna will be others, than for vertical component, but factor sin $(2\pi b/\lambda \sin\theta \cos\beta)$, which is the main thing during the use of a radiation pattern for direction finding, will enter without change and into expression for emf, induced by horizontal field. This fact is utilized in system of two diverse framework (see § 3.7).

In the majority of cases, are applied the antennas, which consist of vertical conductors, free from the reception/procedure of horizontal field. Its remanent/residual reception/procedure is feasible only because of a structural/design or circuit inaccuracy in the production of antennas and coupling feeders. This question is examined in chapter 6, here we will pause at the analysis of formula (3.30) with the reception/procedure of the vertically polarized field.

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Let us assume that each of the antennas is single vertical vibrator (Fig. 3.16). Its radiation pattern does not depend on angle θ and $F_a(\theta, \beta) = F_1(\beta)$.

The phase of resulting emf differs by 90° from the phase of field at the center of system. The amplitude of induced emf, as can be seen from (3.30), directly proportional to the effective height of each of the antennas and depends, furthermore, from angle θ between the direction of propagation of wave and standard to plane AB, the containing antenna, and also from distance between antennas 2b. Radiation patterns in the horizontal plane examined system are represented in Fig. 3.17 for the different designating on curved values $2b/\lambda$. The character of the dependence of resulting emf on angle θ changes during the transition of value $2b/\lambda$ through value of 1/2. Thus far $2b/\lambda < 1/2$, i.e., thus far the distance between antennas is less than the half of wavelength, during a change in the angle θ, are detected two maximums of value E: at $\theta = 90°$ and $\theta = 270°$ and two minimums: at $\theta = 0$ and $\theta = 180°$. This is evident from Fig. 3.17, where are represented diagrams for $2b/\lambda = 1/10$, $2b/\lambda = 1/4$ and $2b/\lambda = 1/2$. The value of maximum emf is determined by the relationship/ratio

$$E_{\text{макс}} = 2Eh_eF_1(\beta)\sin\frac{2\pi b}{\lambda}.$$

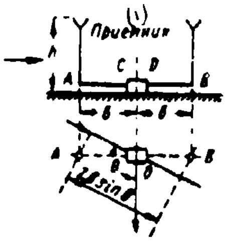

Fig 3 16

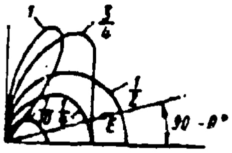

Fig 3 17

Fig. 3.16. Reception/procedure to two spaced antennas.

Key: (1). Receiver.

Fig. 3.17. Radiation patterns of reception/procedure to two spaced antennas.

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This value grow/rises with increase 2b/λ up to value 2b/λ = 1/2, when we obtain a maximally possible value of emf:

$$E_{\text{макс макс}} = 2Eh_cF_1(\beta). \qquad (3.31)$$

During further increase 2b/λ, the value of emf for directions 90 and 270° begins to decrease and the very character of diagram changes: the maximum of emf is obtained not with two, while at four values of angle θ (see Fig. 3.17, radiation patterns for 2b/λ = 3/4 and 2b/λ = 1) determined from the equation:

$$\theta = \arcsin \frac{\lambda}{4b}.$$

With a further increase 2b/λ, the radiation pattern acquires multilobed character.

The value of maximum, as before, is equal to $E_{\text{макс макс}} = 2Eh_e F_1(\beta)$. The minimums of emf and in this case retain their position at θ = 0 and 180°.

Let us pause at the frequently encountered in practice case when 2b/λ << 1. In this case value 2πb/λ sin θ is very low and its sine can be replaced by argument. Then we obtain the following expression for emf of the system:

$$E = \frac{4\pi b h_e}{\lambda} F_1(\beta) E \sin\theta \cos\beta = E_{\text{макс}} \sin\theta\ F_1(\beta) \cos\beta.$$

Radiation pattern in horizontal plane has a form of eight. In vertical plane with $F_1(\beta) = \cos\beta$ (short vibrator) diagram $F(\beta) = \cos^2\beta$.

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Calling the effective height of system h_r the height/altitude of the dipole in which is induced emf, equal to maximum emf, induced in the system of two antennas, we obtain

$$\overset{(1)}{\text{при}}\ \frac{2b}{\lambda} < \frac{1}{2} \quad h_0 = 2h_c \sin\frac{2\pi b}{\lambda},$$

$$\overset{(1)}{\text{при}}\ \frac{2b}{\lambda} \geqslant \frac{1}{2} \quad h_0 = 2h_c.$$

Key: (1). with.

Let us determine the slope/transconductance of the radiation pattern near from the direction of zero reception/procedure, which frequently characterizes the quality of the radio direction finder:

$$F'(0) = \frac{2\pi b}{\lambda} \cos\beta F_1(\beta).$$

Slope/transconductance increases proportional to the distance between antennas. However, with $2b/\lambda > 1/2$, as it is already noted, radiation pattern becomes multilobed, which leads to the multiformity of bearing. The analysis of system from the directional spaced antennas is given into § 3.11.

3.7. System of two diverse framework.

Antenna system consists of two identical framework, been connected in series, one towards by another, and placed on certain distance one from another. Both framework are fastened on the axis, passing through the line of the symmetry of system, and they can together rotate around this axis. Are possible two versions of the execution of the antenna system: the planes of the framework can be or are parallel to the plane, which contains their axes (longitudinal framework, Fig. 3.18a), or are perpendicular to it (transverse framework, Fig. 3.18b).

Under the influence of a normal-polarized electromagnetic wave of emf, induced within each framework, it will be: in the case of the longitudinal framework

$$\dot{E}_{\bullet} = -j\frac{2\pi SN}{\lambda}\mathrm{E}\sin\theta,$$

in the case of the transverse framework

$$\dot{E}_{\bullet} = -j\frac{2\pi SN}{\lambda}\mathrm{E}\cos\theta,$$

where θ is an angle between standard to the line of the connecting framework and direction of propagation.

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Resulting emf in accordance with (3.30) will be obtained equal to:

$$\dot{E} = \frac{4\pi SN}{\lambda}\mathrm{E}\sin\theta\sin\left(\frac{2\pi b}{\lambda}\sin\theta\cos\beta\right) \text{ для } \overset{(1)}{\text{продольных}} \text{ рамок,}$$

$$\dot{E} = \frac{4\pi SN}{\lambda}\mathrm{E}\cos\theta\sin\left(\frac{2\pi b}{\lambda}\sin\theta\cos\beta\right) \text{ для } \overset{(2)}{\text{поперечных}} \text{ рамок.}$$

Key: (1). for the longitudinal framework. (2). for the transverse framework.

If the size/dimensions of the framework and the distance between them are small relative to wavelength, both formulas are simplified. Then for the longitudinal framework

$$\dot{E} = E\,\frac{4\pi^2 SN\,(2b)}{\lambda^2}\sin^2\theta\cos\beta, \qquad (3.32)$$

for the transverse framework

$$\dot{E} = E\,\frac{2\pi^2 SN\,(2b)}{\lambda^2}\sin 2\theta\cos\beta. \qquad (3.33)$$

The radiation patterns, which correspond to formulas (3.32) and (3.33), are represented in Fig. 3.19.

The radiation pattern of the transverse framework has two supplementary of zero. However, this is not created difficulties in the determination of bearing, since with the direction finding of sky waves supplementary zero are obtained ill-defined.

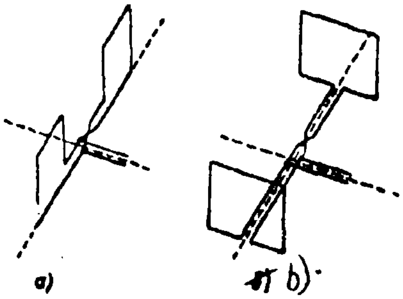

Fig. 3.18. Diverse framework: a) longitudinal; b) transverse.

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Let us examine the effect abnormal-polarized of the component of electric intensity.

If emf, induced by abnormal-polarized field component in each of the framework, is equal to E_r, then resulting emf will be

$$\dot{E}_{r\,\text{рез}} = 2jE_r \sin\left(\frac{2\pi b}{\lambda}\sin\theta\cos\beta\right). \qquad (3.34)$$

In accordance with formula (3.20) during γ = 90° and $\varphi_0 = 0$

$$\dot{E}_r = j\frac{2\pi SN}{\lambda}E\sin\beta\cos\theta \text{ для } \overset{(1)}{\text{продольных}} \text{ рамок,}$$

$$\dot{E}_r = j\frac{2\pi SN}{\lambda}E\sin\beta\sin\theta \text{ для } \overset{(2)}{\text{поперечных}} \text{ рамок.}$$

Key : (1). for the longitudinal framework. (2). for the transverse framework.

From expression (3.34) it is evident that emf, induced by the horizontal component of field, is turned into zero at the same angle θ = 180° at which also it is turned into zero emf, induced by vertical field component. Hence it follows that, although the radiation pattern of the reception/procedure of an abnormal-polarized field differs from the diagram of the reception/procedure of a normal-polarized field, one of the zero directions (perpendicular line, that connects both framework) is retained constant/invariable. This fact determines the possibility of error-less direction finding of sky waves.

From the viewpoint of the possibility of the direction finding of sky waves, which especially steeply fall, is substantial also that that for a vertical electric field the framework does not possess directivity in vertical plane.

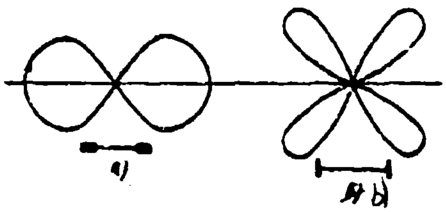

Fig. 3.19. Radiation patterns: a) longitudinal framework; b) transverse framework.

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Therefore the reception/procedure of the steeply incident waves occurs without the weakening of the effect of vertical field component.

In the given analysis are not taken into account the dipoles, equivalent to the action of nonuniform current distribution according to the perimeter of the framework. The circuit diagram of the framework must eliminate reception/procedure to these dipoles in order to avoid the displacement of the direction of zero reception/procedure into the resulting radiation pattern. With the transverse framework (Fig. 3.20) dipoles are included contrarily and are symmetrical earth referenced. Their effect on the displacement of zero directions is eliminated. With the longitudinal framework is possible their location in accordance with Fig. 3.21a or Fig. 3.21b. With location on Fig. 3.21a the equivalent doublets are included to towards each other, but their symmetry is disrupted by the dissimilarity of location relative to feeder. With the location of the framework on Fig. 3.21b the equivalent doublets are vertical and included so that their emf store/add up. Emf of the equivalent doublets will cause the considerable displacement of the minimums of radiation pattern and, therefore, large errors. Thus, one should prefer the use of the transverse framework, presented in Fig. 3.20.

For the exception/elimination of the dependence of the parameters of the framework on the dissimilarity of ground conductivity, and also on the inequalities of soil and for the exception/elimination of antenna effect it is necessary to apply the shielded framework.

Decrease to a certain extent of the effect of the asymmetric location of the framework is achieved during the application/use of the so-called doubled framework (Fig. 3.22).

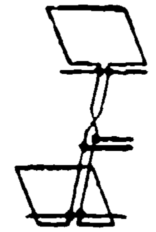

Fig 3.20

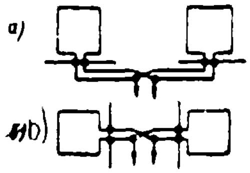

Fig 3.21

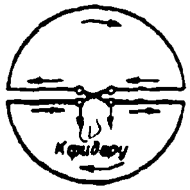

Fig 3.22

Fig. 3.20. Transverse framework with dipoles.

Fig. 3.21. Versions of connection/inclusion of longitudinal framework.

Fig. 3.22. Doubled framework.

Key: (1). To feedar.

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These framework are attachhed to axis in the center of gravity, thanks to which is improved also the mechanical of stability and decreases the moment of the inertia of system during its rotation. Within the doubled framework decreases the harmful effect of the action of the equivalent doublets, since instead of one dipole in them is obtained vapor, operating differentially.

Such framework are of interest, also, for single application/use aboard ship [10.3].

3.8. Combined reception/procedure to the open antenna and the directed system.

Besides reception/procedure to the framework in radio direction finders, is applied the combined reception/procedure to the open antenna and the framework.

Is feasible also the combined reception/procedure to the open antennas and two spaced antennas, or to the open antenna and to goniometric system. theory presented below of the combined reception/procedure is equally used to all these cases.

The schematic of equipment/device with rotatable loop is represented in Fig. 3.23.

Electromagnetic wave excites emf both within the framework and in the open antenna. The appearing in the latter current in turn, induces emf in the duct of the framework because of mutual inductance M. On the grid of receiving tube, operate, thus, two emf.

Emf, induced in the framework by incoming electromagnetic field, according to formula (3.24) can be determined by the expression

$$\dot{E}_р = -jE\frac{2\pi SN}{\lambda}\sin\theta = -jEh_р\sin\theta, \qquad (3.35)$$

where $h_р$ is the affective height of the framework.

Emf, induced in antenna, is found in phase with the field

$$\dot{E}_а = Eh_а.$$

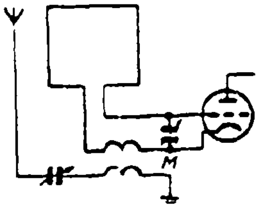

Fig. 3.23. Diagram of the combined reception/procedure to the framework and the antenna.

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Current in the antenna

$$I_a = \frac{E_a}{z_a} e^{j\varphi}, \qquad (3.36)$$

where $$z_a = \sqrt{R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2}, \quad \operatorname{tg}\varphi = \frac{\omega L - \frac{1}{\omega C}}{R}, \qquad (3.37)$$

h_a · the effective height of antenna;

R, ωL, 1/ωC - active and reactance by the antenna of circuit.

During conclusion in view of the smallness of communication/connection the reaction of the framework to antenna we disregard. Emf, induced by this current within the framework, will be

$$E'_p = j\omega M I_a = j\omega M \frac{E_a}{z_a} e^{j\varphi}. \qquad (3.38)$$

Finally, resulting emf within the framework from equations (3.35) and (3.38) will be

$$\dot{E} = \dot{E}_{p} + \dot{E}'_{p} = -jE\left[h_{p}\sin\theta - \omega M\frac{h_{a}}{z_{a}}e^{j\varphi}\right] =$$

$$= -jE\sqrt{\left(h_{p}\sin\theta - \omega M\frac{h_{a}}{z_{a}}\cos\varphi\right)^{2} + \left(\frac{\omega M}{z_{a}}h_{a}\sin\varphi\right)^{2}}\,e^{j\varphi_{1}}, \tag{3.39}$$

where $$\operatorname{tg}\varphi_{1} = \frac{\omega M\frac{h_{a}}{z_{a}}\sin\varphi}{h_{p}\sin\theta - \omega M\frac{h_{a}}{z_{a}}\cos\varphi}.$$

From expression (3.39) it is evident that emf of the framework consists of three terms:

1) emf, induced directly within the framework and dependences on the angle of incident wave θ;

2) by emf, which occurs from the phase is component/term current in antenna and not depending on angle θ; it is found in phase from the first;

3) by emf, which occurs from the extra-phase component of current in antenna and also that which not depends from the angle of incident wave; it is out of phase to angle $\pi/2$ with respect to first emf.

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For larger clarity let us examine two special cases:

1) emf, supplied by antenna, is found in phase from emf, induced in the framework directly, i.e., $\phi = 0$. This case occurs, as can be seen from equation (3.37), when antenna is accurately inclined to incident wave, and therefore $z_{\text{а}} = R$. In this case, resulting emf is equal to

$$\dot{E} = -j\mathrm{E}\left(h_{\text{р}}\sin\theta - \frac{\omega M}{R}h_{\text{а}}\right) = -j\mathrm{E}h_{\text{р}}(\sin\theta - a), \qquad (3.40)$$

where $a = \frac{\omega M}{R}\frac{h_{\text{а}}}{h_{\text{р}}}$ the ratio of emf, induced within the framework from the open antenna, to emf, induced within the framework is direct.

Maximum the amplitude of resulting emf reaches, when $\sin\theta = -1$, i.e., $\theta = 270°$,

$$E_{\text{макс}} = \mathrm{E}h_{\text{р}}(1 + a).$$

At the angle of incident wave $\theta = \arcsin a$, resulting emf is equal to zero. It is obvious, such directions with $a < 1$ will be two, symmetrical relative to the direction oriented radio station.

The derived relationship/ratios are represented graphic on Fig. 3.24. In the form of two concerning circumferences, is represented heart-shaped diagram to the framework, and, since during the transition through 0° emf is changed its sign, right circumference it is marked by sign "+", and left "-". Heart-shaped diagram to antenna is depicted as circumference with center in pole (radius-vector = const). To voltage from antenna is conditionally ascribed sign "+".

Store/adding up taking into account sign in each direction the radius-vector of direct reception in framework and radius-vector of reception through the antenna, we obtain the resulting diagram.

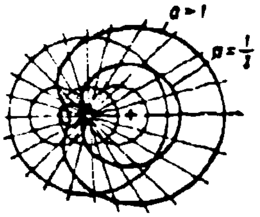

Fig. 3.24. Radiation pattern of the combined reception to the framework and the antenna (case of phase coincidence).

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Addition is made with a = 1/2 and a > 1.

On Fig. 3.25 this adjustment is made for case of a = 1. The obtained in this case curve he is called cardioid. In this case

$$\dot{E} = -j\dot{E}h_{\text{p}}(\sin\theta - 1).$$

Both directions of the zero reception of the combinal diagram are poured into one at θ = 90°.

With a > 1 resulting emf, non-vanishing at which value θ (Fig. 3.24).

Than less by a, the nearer the form of the diagram of reception to eight; than by a more, the more the form of diagram it recalls circle.

2. Emf, induced by antenna, is out of phase in $\pi/2$ relative to its own frame emf, i.e., $\phi = 90°$. This case occurs when antenna is strongly detuned in frequency, since in this case

$$\omega L - \frac{1}{\omega C} \gg R$$

and tg ϕ is great.

To Fig. 3.26, is shown the addition of the directal characteristics for this case. Radius-vectors here must store/add up geometrically, since between them is phase displacement 90°. The module/modulus of resulting emf according to equation (3.39) is equal to

$$E = \mathrm{E}\sqrt{h_p^2 \sin^2\theta + \left(\frac{\omega M}{z_n} h_a\right)^2} = \mathrm{E} h_p \sqrt{\sin^2\theta + a^2}. \qquad (3.41)$$

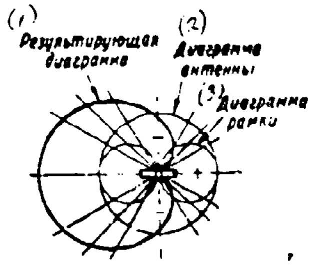

Fig. 3.25.

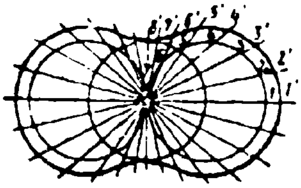

Fig. 3.26.

Fig. 3.25. Cardioid.

Key: (1). Resulting diagram. (2). Diagram of antenna. (3). Diagram of the framework.

Fig. 3.26. Radiation pattern of the combined reception to the framework and the antenna (case of phase displacement in $\pi/2$).

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The maximum of resulting emf occurs at $\theta = 90°$ or $\theta = 270°$:

$$E_{\text{макс}} = Eh_{\text{р}} \sqrt{1+a^2}. \qquad (3.42)$$

Emf is turned here into zero not at which value θ, but has only a minimum with sin $\theta = 0$, i.e., at $\theta = 0°$ and $\theta = 180°$.

The diagram of reception, which is obtained in this case, shows that the conditions of radio traffic deteriorate. Therefore such radiation patterns are undesirable for purposes of direction finding, but, as will be stated in §4.2., they appear with incorrectly selected diagram and construction of external device (presence of the antenna effects).

In the general case when radiation current has two components: in phase and out of phase to 90° relative to emf, are observed the diffuse minimums, which are distinguished between themselves to the angle, unequal 180°.

In the case of the diffuse minimum with direction finding for audition, is observed the angle of the equal to audibility on bisector of which is counted off the bearing.

3.9. Motionless directional antennas with cosinusoidal directional characteristic.

Earlier we examined the directional antennas with cosinusoidal characteristic, which for obtaining a change in the intensity of reception in accordance with directional characteristic must be revolved.

In radio direction finders are applied also the motionless directional antennas. Such antennas can be undertaken greater size/dimensions than rotated, and removed from receiving indicator, which is sometimes necessary (for example, aboard ship). Most frequently is utilized goniometric system.

The operating principle of this system, which consists of two mutually perpendicular framework or the pairs of the spaced antennas, is described into §2.3.

On Fig. 3.27a, is depicted the schematic diagram of goniometric system of two mutually perpendicular framework; on Fig. 3.27b, - a form of the same system in plan/layout.

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Let us examine in more detail processes in system of two framework. The equivalent diagram of this system is given on Fig. 3.28. Let us accept the following designations:

E_1 - emf, induced within the first framework;

E_2 is emf, induced in the second framework;

Z_{11}, Z_{22}, Z_{33} - impedances of the ducts of the first and second framework and search coil of goniometer;

Z_{12} - mutual impedance between the ducts of the framework (with the field coils);

Z_{13}, Z_{22} are mutual impedances between the ducts of each of the framework and the duct of search coil of goniometer;

I_1, I_2, I_3 - point in the ducts of framework and search coil of goniometer.

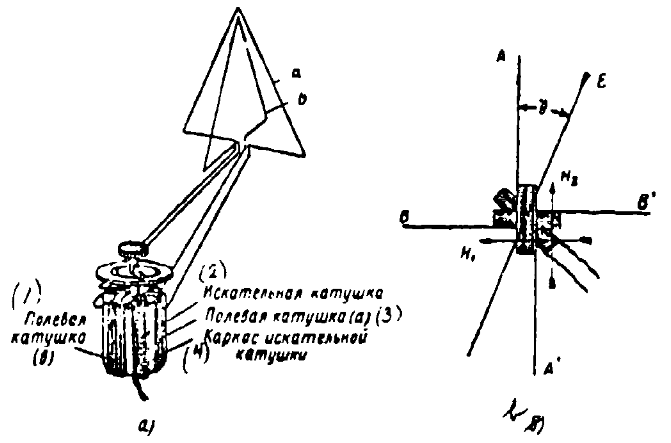

Fig. 3.27. Goniometric system of two mutually perpendicular frameworks: a) general view; b) plan view.

Key: (1). Field coil (c). (2). Search coil. (3). Field coil (a). (4). Framework/body of search coil.

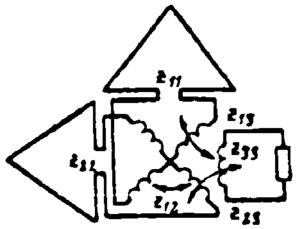

Fig. 3.28. Diagram of goniometric system.

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It is possible to write the following equations:

$$\left.\begin{aligned} I_1 Z_{11} + I_2 Z_{12} + I_3 Z_{13} &= \dot{E}_1, \\ I_1 Z_{12} + I_2 Z_{22} + I_3 Z_{23} &= \dot{E}_2, \\ I_1 Z_{13} + I_2 Z_{23} + I_3 Z_{33} &= 0. \end{aligned}\right\} \qquad (3.43)$$

Solving these equations, we obtain for current I_3 in search coil of the goniometer

$$I_3 = \frac{\dot{E}_1 (Z_{12} Z_{23} - Z_{13} Z_{22}) + \dot{E}_2 (Z_{12} Z_{13} - Z_{11} Z_{23})}{Z_{11} Z_{22} Z_{33} + 2 Z_{12} Z_{13} Z_{23} - Z_{11} Z_{23}^2 - Z_{22} Z_{13}^2 - Z_{33} Z_{12}^2}. \qquad (3.44)$$

We assume that the geometric dimensions of system are small relative to wavelength. Then

$$\left.\begin{aligned} E_1 &= E_{1\,\text{макс}} \cos\theta, \\ E_2 &= E_{2\,\text{макс}} \sin\theta, \end{aligned}\right\} \qquad (3.45)$$

where $E_{1\text{макс}}, E_{2\text{макс}}$ maximum emf within the framework, induced during the coincidence of the plane of the corresponding framework with the direction of incident wave.

Further with the correctly constructed and prepared framework and the goniometer, must be fulfilled the following requirements:

$$\left.\begin{array}{ll} Z_{11}=Z_{22}=Z, & Z_{23}=j\omega M_{2\,\text{макс}}\cos\alpha, \\ Z_{12}=0, & M_{1\,\text{макс}}=M_{2\,\text{макс}}=M, \\ Z_{13}=-j\omega M_{1\,\text{макс}}\sin\alpha, & E_{1\,\text{макс}}=E_{2\,\text{макс}}=Eh_{\text{дф}}, \end{array}\right\} \quad (3.46)$$

where $h_{\text{дф}}$ - the effective height of the framework, in reference to the points of the connection of the field coil of goniometer;

$M_{1\text{макс}}$, $M_{2\text{макс}}$ maximum mutual inductance between search coil and the corresponding field coil of goniometer;

α - the angle between the standard to the first field coil and search coil.

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After substituting expressions (3.45) and (3.46) in (3.44), we will obtain

$$I_3=-j\frac{\omega MEh_{\text{дф}}}{ZZ_{33}-\omega^2M^2}\sin(\theta-\alpha)=$$

$$=-j\frac{\omega MEh_{\text{дф}}}{Z\left(Z_{33}-\frac{\omega^2M^2}{Z}\right)}\sin(\theta-\alpha), \qquad (3.47)$$

where ω^2M^2/Z is the resistor/resistance, introduced by the duct of the framework into the circuit of search coil of goniometer.

Accurately the same current I_2 we would be obtained in the duct, inductively connected with the rotatable loop, turned relative to initial direction in angle α. In this case, α must be equal to the angle of rotation of search coil of goniometer. The parameters of duct and rotatable loop coincide with the parameters of the coils of goniometer and framework of goniometric system, but the mutual inductance of the duct and framework is equal to the maximum mutual inductance M of the searching and field coils of goniometer. The equivalent diagram of system with two framework (Fig. 3.23) let us be guided during the calculation of the effectiveness of goniometric system (see Chapter 7).

The obtained results can be common for goniometric system of two pairs of the spaced antennas.

The coils of goniometer it is possible to wind on framework/body from insulation (air goniometer). Coupling coefficient between field and search coils of air goniometer is limited to value 0.4-0.5.

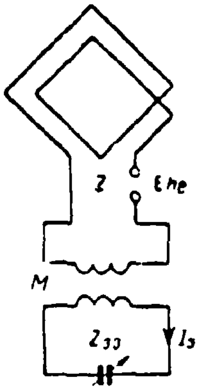

Fig. 3.29. Equivalent diagram of goniometric system.

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With the target/purpose of an increase in the coupling coefficient, under the condition of the uniformity of magnetic flux, are applied also the goniometers with ferromagnetic cores (Fig. 3.30). As material are applied the magnetic dielectrics (carbonyl, Alsifer and others) and ferrites.

In such goniometers usually the rotor (search coil) is placed within stator (field) coils. On medium-frequency waves there are constructions where the stator coils are mounted within rotor coil

[10.5].

Of ferrite core goniometers, the coupling coefficient between searching and field coils reaches 0.7-0.95. Good accuracy with large coupling coefficient provide coils with uniform winding/coil on ferrite tori. Beginning and the end/lead of each winding are connected. Removal/outlets of stator coil for the connection of the framework are taken through 90°. Of rotor the removal/outlets are taken from the diametrically opposite points of winding. For obtaining more uniform field, it is necessary to carry out a nonuniform winding/coil of rotor, approximately according to sinusoidal law, and to take removal/outlets from the middles of the most diverse turns. Sometimes of rotor are made the supplementary removal/outlets, shifted to 90° relative to fundamental, for use during the determination of the side of radio station [10.5, 4.6]. For a decrease in the capacitive coupling, is applied the electrostatic shield (see Fig. 4.17). However, an increase in the capacitance of windings by housing impedes the use of such goniometers at more high frequencies (large 25-30 MHz).

Instead of the inductive goniometer it is possible to apply the capacitive goniometer, which is adjustable capacitor of two systems of stator plates and one rotor.

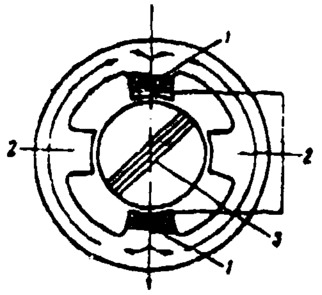

Fig. 3.30. Goniometer with the ferromagnetic core: 1 - field coil; 2 - field coil; 3 - search coil.

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If is made the requirement for the cosinusoidal law of a change in the communication/connection between the rotor and each stator, which is realized with the special form of the plates of rotor, then for a capacitive goniometer they remain valid of the relationship/ratios, derived for an inductive goniometer.

To Fig. 3.31, is shown the schematic diagram of the input part of the goniometric radio direction finder with capacitive goniometer during the application/use of spiral loops. In the used at present goniometric systems, predominantly with inductive goniometer, are utilized the unadjusted framework. The application/use of spiral loops is inconvenient, since during their use is required the preliminary fine tuning of both framework for the frequency of the oriented radio station on the local oscillator, which complicates work [4.8, 8.29].

3.10. Goniometric system from n of the spaced antennas.

In §3.9 is examined antenna system of four spaced antennas. In the radio direction finder of goniometric system with the spaced vertical wire antennas, it is possible to use another number of antennas. Let us examine the general case by the antenna of system with n by the spaced vertical wire antennas (Fig. 3.32). Antennas are arrange/located in circumference at equal angular distance 2π/n one from another (1, 2, 3, ..., n), the radius, carried out to the n antenna, coinciding with initial reference line. Each antenna is connected to the appropriate field coil of goniometer (I, II, III, ..., N). The number of field coils is equal to the number of antennas: N = n. The mutual location of field coils on common/general/total framework corresponds to three-dimensional/space antenna location. Field coils are connected by star, common point can be grounded. Inside of field coils rotates search coil (bb on Fig. 3.32a).

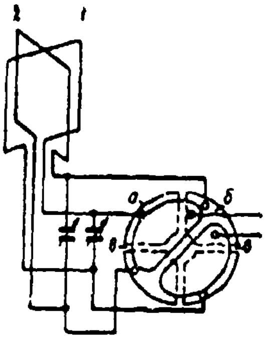

Fig. 3.31. The schematic diagram of goniometric radio direction finder with the capacitive goniometer: 1 - the first framework; a) plate the 1st stator; b) the plate of the 2nd stator; c) the plate of rotor; 2. the second framework.

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Each field coil creates its magnetic field. If we accumulate the magnetic fields of all field coils, then in the correctly designed radio direction finder, as in the simplest goniometric system with two mutually perpendicular framework, the direction of the resulting field in goniometer forms with standard to the n field coil the same angle, which the direction of the oriented radio station is formed with initial reference line.

According to the scale of goniometer with auditory direction-finding method, is counted off the angle between the plane of search coil and the perpendicular to the n field coil. When current strength in search coil is equal to zero, This angle corresponds to bearing to radio station.

Let us determine current in search coil of goniometer.

Let us designate:

E_0 - the strength of field at the center of system;

θ is an angle of the direction of the arrival of wave in horizontal plane with the radius, passing through n-s antenna;

β - the angle of the slope of a front of wave;

2b - the diameter of a circle of the arrangement/permutation of antennas.

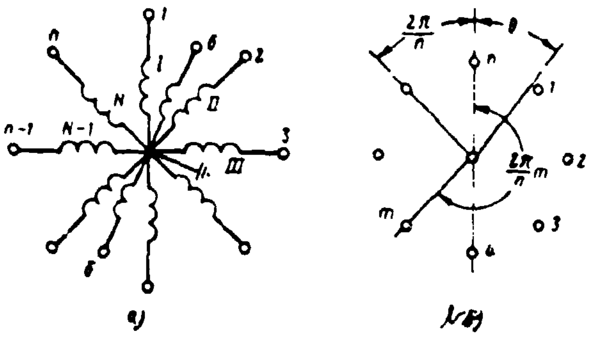

Fig. 3.32. Antenna system from n of the spaced antennas: a) the circuit of the connection of the field coils of goniometer (1, 2, 3, ..., n - antenna; I, II, III, ..., N - the field coils of goniometer, b-b - search coil of goniometer); b) location n of antennas in plan/layout.

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On n-th antenna operates the electric field with strength $E_n = E_0 e^{j\frac{2\pi R}{\lambda}\cos\beta\cos\theta}$.

At the point of the location of the m th antenna, distant to angle 2πm/n from the n-th, the strength of field will be where $a = 2\pi b/\lambda \cos \beta$, $\delta = 2\pi/n$, $\theta - \delta m = \gamma$.

It is known that $e^{ja\cos\gamma}$ it is possible to expand in Fourier series - Bessel

$$e^{ja\cos\gamma} = J_0(a) + 2\sum_{p=1}^{\infty} j^p J_p(a)\cos p\gamma, \qquad (3.49)$$

and is analogous

$$e^{j\frac{2\pi b}{\lambda}\cos\beta\cos\left(\theta-\frac{2\pi m}{\lambda}\right)} = J_0\left(\frac{2\pi b}{\lambda}\cos\beta\right) + $$
$$+ 2\sum_{p=1}^{\infty} j^p J_p\left(\frac{2\pi b}{\lambda}\cos\beta\right)\cos p\left(\theta - \frac{2\pi m}{n}\right), \qquad (3.49')$$

where J_0, J_1, J_2, ..., J_p with first-order Bessel function the zero, first, second, ..., the p-th of orders.

If we designate by h_{cn} the effective height of antenna, then emf of the m antenna will be

$$\dot{E}_m = E_0 h_{co}\left[J_0\left(\frac{2\pi b}{\lambda}\cos\beta\right) + \right.$$
$$\left. + 2\sum_{p=1}^{\infty} j^p J_p\left(\frac{2\pi b}{\lambda}\cos\beta\right)\cos p\left(\theta - \frac{2\pi m}{n}\right)\right] =$$
$$= \sum_{p=0}^{\infty} A_p \cos p\left(\theta - \frac{2\pi m}{n}\right) = \sum_{p=0}^{\infty} A_p \cos p\theta \cos\frac{2\pi}{n} pm +$$
$$+ \sum_{p=0}^{\infty} A_p \sin p\theta \sin\frac{2\pi}{n} pm, \qquad (3.50)$$

where

$$A_0 = E_0 h_{e0} J_0 \left(\frac{2\pi h}{\lambda}\cos\beta\right),\quad A_{p(p\geq 1)} = 2E_0 h_{e0} j^p J_p \left(\frac{2\pi h}{\lambda}\cos\beta\right). \tag{3.51}$$

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Equivalent diagram for calculation.

For determining current in search coil of goniometer, one should solve system from (n + 1) the equations:

$$\left.\begin{array}{l} I_1 Z_{11} + I_2 Z_{12} + I_3 Z_{13} + \ldots + I_n Z_{1n} + I_{\text{н}} Z_{1\text{н}} = E_1, \\ I_1 Z_{21} + I_2 Z_{22} + I_3 Z_{23} + \ldots + I_n Z_{2n} + I_{\text{н}} Z_{2\text{н}} = E_2, \\ \cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots \\ \cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots \\ I_1 Z_{n1} + I_2 Z_{n2} + I_3 Z_{n3} + \ldots + I_n Z_{nn} + I_{\text{н}} Z_{n\text{н}} = E_n, \\ I_1 Z_{\text{н}1} + I_2 Z_{\text{н}2} + I_3 Z_{\text{н}3} + \ldots + I_n Z_{\text{н}n} + I_{\text{н}} Z_{\text{нн}} = 0. \end{array}\right\} \tag{3.52}$$

Here I_1, I_2, I_3, ..., I_n — currents in the field coils of goniometer [1];

Z_{11}, Z_{22}, Z_{33}, ..., Z_{nn} - the internal resistances of the circuits of field coils with the appropriate antennas;

I_{nn} and Z_{nn} - current and the resistor/resistance of the duct of search coil of goniometer;

$Z_{12}, Z_{13}, \ldots, Z_{mj}$ - mutual impedances of circuits the 1st and 2nd, 1st and 3rd, m-th and j-th field coils;

$Z_{\text{н}1}, Z_{\text{н}2}, \ldots, Z_{\text{н}n}$ mutual impedances of field coils with search coil of goniometer.

FOOTNOTE [1]. We consider that the currents in antenna mounting and in the field coil of goniometer coincide in value and in phase. Calculations are given in chapter 7. ENDFOOTNOTE.

We consider that network elements of all antennas are completely identical.

Then

$$Z_{11} = Z_{22} = \ldots = Z_{mm} = \ldots = Z_{nn} = \ldots = Z_{\text{а}0} + j\omega L_{\text{н}}, \tag{3.53}$$

where $Z_{\text{а}0}$ the resistor/resistance of antenna itself;

$L_{\text{н}}$ the inductance of field coil.

On the strength of the symmetry of the arrangement/permutation of antennas (see Fig. 3.32b)

$$Z_{m,m+1}=Z_{m,m-1}=Z_{c1};\ Z_{m,m+2}=Z_{m,m-2}= \\ =Z_{c2};\ \ldots\ Z_{m,m+k}=Z_{m,m-k}=Z_{ck}. \qquad (3.54)$$

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These resistor/resistances are accumulated from mutual impedance of antennas themselves and from mutual inductance between the field coils, connected in these antennas.

The solution of system of equations (3.52) for all currents very cumbersomely and is not given.

Without solving system of equations (3.52), it is possible to conduct the calculation of current in search coil in the following order:

a) to determine currents in each field coil, on the basis of emf of antenna and impedance of antenna circuit, taking into account the resistor/resistances, introduced from the circuits of other antennas, at the broken circuit of search coil of the goniometer when it not coupled impedance into antenna circuit;

b) to find the open-circuit voltages, induced by the currents of field coils in search coil of goniometer;

c) to calculate impedance of the duct of search coil taking into account the resistor/resistances, introduced from the circuits of all antennas;

d) to determine current in the duct of search coil of goniometer.

It is feasible calculations in the indicated order.

a) for the calculation of the current of the m^{th} antenna we will use equation from system (3.52) for the circuit of the m^{th} antenna:

$$I_1 Z_{m1} + I_2 Z_{m2} + \ldots + I_m Z_{mm} + \ldots + I_n Z_{mn} + \\ + I_N Z_{mN} = \dot{E}_m. \qquad (3.55)$$

We assume that because of the symmetry of the system of the antennas of the amplitude of currents in all antenna circuits are identical, the phases of currents are determined by the phases of emf in the appropriate antennas, i.e.,

$$\left.\begin{aligned} I_1 &= I_{\text{макс}} e^{ja\cos(\theta-\delta)}, \\ I_2 &= I_{\text{макс}} e^{ja\cos(\theta-2\delta)}, \\ &\cdots\cdots\cdots\cdots \\ I_m &= I_{\text{макс}} e^{ja\cos(\theta-m\delta)}, \\ &\cdots\cdots\cdots\cdots \\ &\cdots\cdots\cdots\cdots \\ I_n &= I_{\text{макс}} e^{ja\cos\theta}. \end{aligned}\right\} \tag{3.56}$$

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After substituting in (3.55) expressions (3.48), (3.53), (3.54) and (3.56), we will obtain

$$\begin{gathered} I_{\text{макс}}\{(Z_{\text{п0}} + j\omega L_{\text{п}})\, e^{ja\cos(\theta-\delta m)} + Z_{\text{c1}}[e^{ja\cos[\theta-\delta(m-1)]} + \\ + e^{ja\cos[\theta-\delta(m+1)]}] + Z_{\text{c2}}[e^{ja\cos[\theta-\delta(m-2)]} + \\ + e^{ja\cos[\theta-\delta(m+2)]}] + \ldots + Z_{\text{c}\frac{n}{2}}[e^{-ja\cos(\theta-\delta n)}]\} = \\ = \dot{E}_m = E_0 h_{\text{c0}} e^{ja\cos(\theta-\delta m)} \end{gathered} \tag{3.57}$$

for even number of antennas. If the number of antennas is odd, then last/latter term in the curly braces is absent. Let us use the formula of expansion (3.49') for the components of expression (3.57), moreover we will be restricted by two members of series [1]:

$$
\begin{aligned}
I_{\text{макс}}\{(Z_{a0}+j\omega L_{\text{п}})[J_0(a)+j2J_1(a)\cos(\theta-\delta m)]+\\
+2Z_{c1}[J_0(a)+j2J_1(a)\cos(\theta-\delta m)\cos\delta]+\\
+2Z_{c2}[J_0(a)+j2J_1(a)\cos(\theta-\delta m)\cos 2\delta]+\ldots+\\
+Z_{c\frac{n}{2}}[J_0(a)-j2J_1(a)\cos(\theta-\delta m)]\}=\\
=E_0h_{c0}[J_0(a)+j2J_1(a)\cos(\theta-\delta m)]. \qquad (3.58)
\end{aligned}
$$

FOOTNOTE [1]. This limitation sufficiently for the establishment of the conditions of the error-free operation of system (see §4.3). ENDFOOTNOTE.

The voltage of antenna has two components: depending on the direction of the arrival of wave $j2E_0h_{c0}J_1(a)\times\cos(\theta-\delta m)$ and independent of it $E_0h_{c0}J_0(a)$. We respectively have two different equations for determining antenna resistance.

For the component voltage of the antenna, which depends on the direction of the arrival of wave, complete antenna resistance is expressed

$$Z_{a\Sigma}=Z_{a0}+j\omega L_{\text{п}}+\sum_{m=1}^{n-1}Z_{cm}\cos\delta m. \qquad (3.59)$$

independent of direction θ and the number of antenna. For the component voltage of the antenna, independent of the direction of the arrival of wave, the expression for complete antenna resistance takes the form

$$Z'_{a\Sigma} = Z_{a0} + j\omega L_n + \sum_{m=1}^{n-1} Z_{cm}. \qquad (3.59')$$

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Let us calculate antenna resistance (3.59).

Mutual impedances of antenna circuits Z_{cm} consist of mutual impedances of antennas themselves Z_{cmA} and the mutual inductance of the field coils, connected in antennas. We assume that if the angle between the planes of two field coils is equal to δm, then the mutual inductance of such coils is equal to KL_n cos δm, where K is a coupling coefficient between the field coils, planar.

Therefore

$$\sum_{m=1}^{n-1} Z_{cm} \cos \delta m = \sum_{m=1}^{n-1} Z_{cm\,a} \cos \delta m + j\omega L_n K \sum_{m=1}^{n-1} \cos^2 \delta m.$$

Since

$$\sum_{m=1}^{n-1} \cos^2 \delta m = \frac{n}{2} - 1,$$

then

$$\sum_{m=1}^{n-1} Z_{\text{c}m}\cos\delta m = \sum_{m=1}^{n-1} Z_{\text{c}m\,\text{a}}\cos\delta m + j\omega L_{\text{п}}K\left(\frac{n}{2}-1\right)$$

and complete antenna resistance $(Z_{\text{a}\Sigma})$ will be

$$Z_{\text{a}\Sigma} = Z_{\text{a}0} + \sum_{m=1}^{n-1} Z_{\text{c}m\,\text{a}}\cos\delta m + j\omega L_{\text{п}}\left[K\left(\frac{n}{2}-1\right)+1\right]. \tag{3.60}$$

The amplitude of current in antenna is designed from the formula

$$I_{\text{макс}} = \frac{E_0 h_{e0}}{Z_{\text{a}\Sigma}}. \tag{3.61}$$

b) let us determine the open-circuit voltage $E_{\text{п}}$ with the induced currents of antennas I_m in search coil of goniometer, considering that search coil forms angle α with standard to the plane of the n field coil and that the magnetic fields of field coils are uniform.

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Let us designate:

$M_{\text{макс}}$ maximum mutual inductance between field and search coils of goniometer;

$M_{m\text{п}} = M_{\text{макс}}\sin(\delta m - \alpha)$ mutual inductance between mth field and search coils of goniometer.

Then

$$\dot{E}_{\text{и}} = j\omega M_{\text{макс}} \sum_{m=1}^{n} \sin(\delta m - \alpha) I_m = $$
$$= j\frac{\omega M_{\text{макс}}}{Z_{\text{вх}}} \sum_{m=1}^{n} E_m \sin(\delta m - \alpha). \qquad (3.62)$$

After substituting in (3.62) expression (3.50) for E_m, we will obtain

$$\dot{E}_{\text{и}} = j\frac{\omega M_{\text{макс}}}{Z_{\text{вх}}} \sum_{m=1}^{n} \Big[\Big(\sum_{p=0}^{\infty} A_p \cos p\theta \cos \frac{2\pi}{n} pm +$$
$$+ \sum_{p=0}^{\infty} A_p \sin p\theta \sin \frac{2\pi}{n} pm \Big) \sin\Big(\frac{2\pi}{n} m - \alpha\Big)\Big]$$

or

$$\dot{E}_{\text{и}} = j\frac{\omega M_{\text{макс}}}{Z_{\text{вх}}} \sum_{m=1}^{n} \Big\{ -\Big[\Big(\sum_{p=0}^{\infty} A_p \cos p\theta \cos \frac{2\pi}{n} pm \times$$
$$\times \cos\frac{2\pi}{n} m + \sum_{p=1}^{\infty} A_p \sin p\theta \sin \frac{2\pi}{n} pm \cos \frac{2\pi}{n} m \Big) \sin\alpha \Big] +$$
$$+ \Big[\Big(\sum_{p=0}^{\infty} A_p \cos p\theta \cos \frac{2\pi}{n} pm \sin \frac{2\pi}{n} m +$$
$$+ \sum_{p=1}^{\infty} A_p \sin p\theta \sin \frac{2\pi}{n} pm \sin \frac{2\pi}{n} m \Big) \cos\alpha \Big]\Big\}. \qquad (3.63)$$

After using the transformations, given in appendix II, we will obtain:

$$\dot{E}_{н} = j\frac{\omega M_{макс}}{Z_{эт}}\frac{n}{2}\left\{\left[\sum_{k=0}^{\infty} A_{kn+1}\sin(kn+1)\theta - \sum_{k=1}^{\infty} A_{kn-1}\sin(kn-1)\theta\right]\cos\alpha - \left[\sum_{k=0}^{\infty} A_{kn+1}\cos(kn+1)\theta + \sum_{k=1}^{\infty} A_{kn-1}\cos(kn-1)\theta\right]\sin\alpha\right\}. \qquad (3.63')$$

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Let us write expressions for $E_{н}$ in the following form:

$$\dot{E}_{н} = \frac{\omega M_{макс}}{Z_{эт}} H_{аф} E_{0}, \qquad (3.64)$$

where $H_{аф}$ the equivalent effective height for the directional reception of goniometric system from n of the vertical wire antennas.

On the basis (3.63'), (3.51) and (3.64) we have

$$H_{аф} = h_{co} n\left\{J_{1}\left(\frac{2\pi}{\lambda} b\cos\beta\right)\sin(\theta-\alpha) + j^{n-2}J_{n-1}\left(\frac{2\pi}{\lambda} b\cos\beta\right)\sin[(n-1)\theta-\alpha] + j^{n}J_{n+1}\left(\frac{2\pi}{\lambda} b\cos\beta\right)\sin[(n+1)\theta-\alpha] + \ldots\right\}. \qquad (3.65)$$

FOOTNOTE 1. Before the expression $H_{a\varphi}$ is lowered minus sign. Terms with ' designate stress component, out of phase in π/2 from fundamental. For greater detail, see §4.9. ENDFOOTNOTE.

If the diameter of the arrangement/permutation of antennas (separation of antennas) 2b is selected so that it is possible to be restricted one first term of a series (3.65), then

$$H_{a\varphi} = h_{c0} n J_1 \left(\frac{2\pi}{\lambda} b \cos\beta\right) \sin(\theta - \alpha)$$

and

$$\dot{E}_n = \frac{\omega A I_{\text{макс}}}{Z_{a\Sigma}} H_{a\varphi 0} E_0 \sin(\theta - \alpha). \qquad (3.64')$$

where

$$H_{a\varphi 0} = h_{c0} n J_1 \left(\frac{2\pi}{\lambda} b \cos\beta\right). \qquad (3.66)$$

With $\beta = 0$ we have

$$H'_{a\varphi 0} = h_{c0} n J_1 \left(\frac{2\pi}{\lambda} b\right). \qquad (3.66')$$

The equivalent effective height $H'_{a\varphi 0}$ is maximum, when $J_1(2\pi/\lambda\ b)$ = max. First root of this equation - 1.84 or 2b/λ = 0.586. With $J_1(2\pi/\lambda\ b) = 0$ equivalent actual height/altitude is equal to zero.

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The smallest value, which satisfies this condition, will be 2π/λ b = 3.83 or 2b/λ = 1.22.

Finally, in the case when 2b/λ << 1 and J_1 (2π/λ b) ≈ πb/λ, the effective height

$$H''_{\text{аф0}} \approx \frac{2\pi}{\lambda} b h_{\text{со}} \frac{n}{2}. \qquad (3.67)$$

c) Impedance $Z_{\text{п}}$ of search coil consists of inductive reactance of search coil $j\omega L_{\text{п}}$, load impedance $Z_{\text{н}}$ and the resistor/resistances, introduced from the circuits of the field coils of all antennas.

Mutual inductance between a-field ii by search coils

$$M_{mn} = M_{\text{макс}} \sin(\delta m - \alpha).$$

Impedance of search coil has the expression

$$Z_{\text{п}} = j\omega L_{\text{п}} + Z_{\text{п}} + \sum_{m=1}^{n} \frac{\omega^2 M^2_{m\text{п}}}{Z_{\text{а}\Sigma}} = j\omega L_{\text{п}} + Z_{\text{п}} + \\ + \frac{\omega^2 M^2_{\text{макс}}}{Z_{\text{а}\Sigma}} \sum_{m=1}^{n} \sin^2 (\delta m - \alpha).$$

Taking into account that $\delta = 2\pi/n$,

$$\sum_{m=1}^{n} \sin^2 \frac{2\pi}{n} m = \sum_{m=1}^{n} \cos^2 \frac{2\pi}{n} m = \frac{n}{2}$$

and

$$\sum_{m=1}^{n} \sin \frac{2\pi}{n} m = \sum_{m=1}^{n} \cos \frac{2\pi}{n} m = 0,$$

we obtain for $Z_{\text{н}}$

$$Z_{\text{н}} = j\omega L_{\text{н}} + Z_{\text{н}} + \frac{\omega^2 M^2_{\text{макс}}}{Z_{\text{а}\Sigma}} \frac{n}{2}. \qquad (3.68)$$

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d) Expression for a current in search coil on the basis (3.64') and (3.68) with not the very large separation of antennas [see (3.64')] will be

$$I_{\text{п}} = \frac{\dot{E}_{\text{п}}}{Z_{\text{п}}} \approx \frac{\omega M_{\text{макс}}}{Z_{\text{а}\Sigma} Z_{\text{п}}} H_{\text{д}\varphi_0} \text{E}_0 \sin (\theta - \alpha) = \\ = \frac{\omega M_{\text{макс}} h_{\text{д}0} I_1 \left(\frac{2\pi}{\lambda} b \cos \beta \right) n}{(j\omega L_{\text{п}} + Z_{\text{п}}) Z_{\text{а}\Sigma} + \omega^2 M^2_{\text{макс}} \frac{n}{2}} \text{E}_0 \sin (\theta - \alpha). \qquad (3.69)$$

We examine the case when the number of antennas n - even and opposite antennas are connected to one and the same field coil, then the number of field coils N = n/2.

Let us leave for the inductance of field and search coils of goniometer, and also for the mutual inductance of coils the adopted previously designations:

$$L_{\text{п}},\ L_{\text{и}},\ M_{\text{пи}},\ M_{\text{п}m}=K\omega L_{\text{и}}\cos\left(\frac{2\pi m}{n}\right),$$

only m here varies from 1 to N. From formula (3.60) the expression for impedance of the pair of antennas will be

$$\begin{aligned}Z_{\text{a}\Sigma}&=2Z_{\text{a}_0}+j\omega L_{\text{п}}+2\sum_{m=1}^{n-1}Z_{\text{c}m\,\text{a}}\cos\delta m+\\&+j\omega L_{\text{п}}K\sum_{m=1}^{N-1}\cos^2\delta m=2Z_{\text{a}_0}+2\sum_{m=1}^{n-1}Z_{\text{c}m\,\text{a}}\cos\delta m+\\&+j\omega L_{\text{п}}\left[K\left(\frac{N}{2}-1\right)+1\right].\end{aligned}\tag{3.70}$$

With K = 1

$$Z_{\text{вх}}=2Z_{\text{а0}}+2\sum_{m=1}^{n-1} Z_{\text{см а}}\cos\delta m+j\omega L_{\text{п}}\frac{N}{2}. \quad (3.70')$$

The resistor/resistance of search coil we will obtain from (3.68) in the form

$$Z_{\text{н}}=j\omega L_{\text{п}}+Z_{\text{н}}+\frac{\omega^2 M^2_{\text{макс}}}{Z_{\text{аΣ}}}\frac{N}{2}.$$

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Let us relate the equivalent effective height of system $H_{\text{эф0}}$ to the effective height of the pair of antennas, which let us designate

$$H_0=2h_{e0}\sin\left(\frac{2\pi b}{\lambda}\cos\beta\right)\approx 4h_{e0}J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right).$$

Then from (3.66) we obtain that $H_{\text{эф0}}=H_0\frac{N}{2}$.

Current in search coil is designed from the formula

$$I_{\text{п}}=\frac{\omega M_{\text{макс}}H_0\frac{N}{2}E_0}{(j\omega L_{\text{н}}+Z_{\text{п}})Z_{\text{аΣ}}+\omega^2 M^2_{\text{макс}}\frac{N}{2}}\sin(\theta-\alpha). \quad (3.71)$$

When $\theta=\alpha$ $I_{\text{п}}=0$, at $\theta=\alpha+90°$ $I_{\text{п}}=I_{\text{п макс}}$.

On Fig. 3.33, is given equivalent diagram for the calculation of the current of search coil of the goniometer of system from N of the pairs of the vertical wire antennas. Applied emf corresponds by maximum emf of one pair of antennas, multiplied by $\sqrt{\frac{N}{2}}$. In accordance with (3.70) the inductance of the first duct (field coil) $L_{\text{по}}=L_{\text{п}}\left[K\left(\frac{N}{2}-1\right)+1\right]$. Coefficient K can be taken as equal to unity, then $L_{\text{по}}=L_{\text{п}}\frac{N}{2}$. Coupling coefficient of both ducts $\left(K_{\text{э}}=K_{\text{r}}\times\sqrt{\frac{N/2}{KN/2-K+1}}\right.$; with K = 1 $K_{\text{э}}=K_{\text{r}}$ to the coupling coefficient field and search coils.

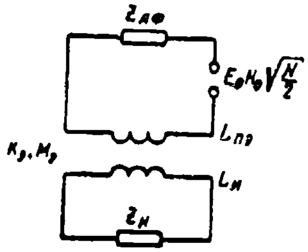

Fig. 3.33. Equivalent diagram for the calculation of goniometric system with n by the spaced antennas.

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Equivalent mutual inductance of ducts $M_\text{э} = M_\text{макс} \sqrt{\frac{N}{2}}$.

Antenna resistance is expressed

$$Z_{a\varphi} = 2Z_{ae} + 2\sum_{m=1}^{n-1} Z_{\text{с}ma} \cos \frac{2\pi m}{n}.$$

Current in search coil of goniometer taking into account equivalent diagram and (3.71) is designed from the formula

$$I_{\text{н макс}} = \frac{\omega M_\text{э} H_\Phi \sqrt{\frac{N}{2}} E_\Phi}{(j\omega L_\text{н} + Z_\text{н})(Z_{a\varphi} + j\omega L_{n\text{э}}) + \omega^2 M_\text{э}^2} = \frac{\omega M_\text{э} H_\Phi \sqrt{\frac{N}{2}} E_\Phi}{(Z_{a\varphi} + j\omega L_{n\text{э}})\left(j\omega L_\text{н} + Z_\text{н} + \frac{\omega^2 M_\text{э}^2}{Z_{a\Phi} + j\omega L_{n\text{э}}}\right)}. \quad (3.72)$$

In two-channel visual radio direction finder (see §8.3) during application/use by n- antenna of system is utilized the matching cell/element, which is goniometer with $N = n/2$ field and with two mutually perpendicular search coils from which are remove/taken the voltages on channels. Structurally cell/element can be made so that search coils do not rotate and have constant orientation with respect to field coils. This matching cell/element he is sometimes called coordinate transformer.

Let us designate the angle, formed by the plane of search coil of the first channel with standard to the n^{th} field coil, $\pi/2 - \alpha$. Second search coil forms with the same standard angle α. Then the currents in the circuits of search coils of coordinate transformer are determined from the formulas

$$I_{н1}=\frac{\omega M_0 H_0 \sqrt{\frac{N}{2}}\,E_0}{(j\omega L_{н1}+Z_{н1})(Z_{а0}+j\omega L_{а0})+\omega^2 M_0^2}\cos(\theta-\alpha),$$

$$I_{н2}=\frac{\omega M_0 H_0 \sqrt{\frac{N}{2}}\,E_0}{(j\omega L_{н2}+Z_{н2})(Z_{а0}+j\omega L_{а0})+\omega^2 M_0^2}\sin(\theta-\alpha)$$

and during satisfaction of conditions (3.96) the relation

$$\frac{I_{н2}}{I_{н1}}=\operatorname{tg}(\theta-\alpha).$$

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System with parallel connection of adjacent antennas.

In radio direction finder with n antennas, it is possible to in parallel connect the adjacent pairs of antennas. The field coils of goniometer are included in this case, as is shown Fig. 3.34. Angle 2γ between the connected in parallel antennas can be not equal to 2π/n. This connection of antennas makes it possible to decrease two times the number of field coils of goniometer and, for example, in eight-antenna system to use goniometer with two field coils. During this connection also is facilitated the connection to eight-antenna systems, which obtained wide application, two-channel receiving indicator.

For this case is retained entire analysis, given earlier, only in the course of calculation one should make the following changes: instead of n to take n/2; instead of each term of expansion (3.63') to take the sum of two terms, in one of which $(k\,n/2 \pm 1)\theta$ is replaced by $(k\,n/2 \pm 1) \times (\theta + \gamma)$, in the other, by $(k\,n/2 \pm 1)(\theta - \gamma)$, where 2γ is an angle between the connected in parallel antennas (Fig. 3.34).

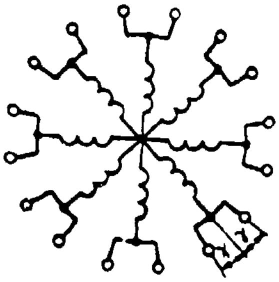

Fig. 3.34. Circuit diagram of field coils during parallel connection of adjacent antennas.

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With this addition we will obtain

$$\dot{E}_{н}=j\,\frac{\omega M_{\text{макс}}}{Z_{\text{вх}}}\,\mathrm{E}_0 h_{e0}\frac{n}{2}\left\{\left[\sum_{k=0}^{\infty}A_{k\frac{n}{2}+1}\sin\left(k\frac{n}{2}+1\right)\theta\times\right.\right.$$
$$\times\cos\left(k\frac{n}{2}+1\right)\gamma-\sum_{k=1}^{\infty}A_{k\frac{n}{2}-1}\sin\left(k\frac{n}{2}-1\right)\theta\times$$
$$\times\cos\left(k\frac{n}{2}-1\right)\gamma\right]\cos\alpha-\left[\sum_{k=0}^{\infty}A_{k\frac{n}{2}+1}\cos\left(k\frac{n}{2}+1\right)\theta\times$$
$$\left.\left.\times\cos\left(k\frac{n}{2}+1\right)\gamma+\sum_{k=1}^{\infty}A_{k\frac{n}{2}-1}\cos\left(k\frac{n}{2}-1\right)\gamma\right]\sin\alpha\right\}. \tag{3.73}$$

As before, when 2b/λ << 1,

$$\dot{E}_{н}=j\,\frac{\omega M_{\text{макс}}}{Z_{\text{вх}}}\,\mathrm{E}_0 h_{e0} n J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right)\cos\gamma\sin(\theta-\alpha)$$ and when $\theta=\alpha$ $E_{н}=0.$

3.11. Antenna systems with acute/sharp directional characteristic.

Application/use n of antennas, straight.

For by the antenna of the system of the directional reception it is possible to utilize n of the identical antennas, straight (Fig. 3.35) and parallel-connected. Let us designate d the distance between adjacent antennas, $2b = d(n - 1)$ - the complete separation between extreme antennas.

It is known that during parallel (cophasal) connection of antennas standardized/normalized directional characteristic is expressed

$$F(\theta, \beta) = F_1(\theta, \beta) \frac{\sin\left(\frac{n}{2} md \sin\theta \cos\beta\right)}{n \sin\left(\frac{1}{2} md \sin\theta \cos\beta\right)}, \quad (3.74)$$

where $F_1(\theta, \beta)$ - directional characteristic of single antenna; θ it is counted off from perpendicular to the line of antennas.

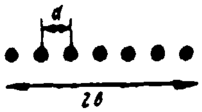

Fig. 3.35. Antenna system from n of the antennas, straight.

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When $\beta = 0$, is obtained radiation pattern in horizontal plane. Then

$$F(\theta) = F_1(\theta) \frac{\sin\left(\frac{n}{2} md \sin\theta\right)}{n \sin\left(\frac{1}{2} md \sin\theta\right)}. \quad (3.75)$$

Examples of the radiation patterns of the group of the vertical wire antennas in horizontal plane are given on Fig. 3.36.

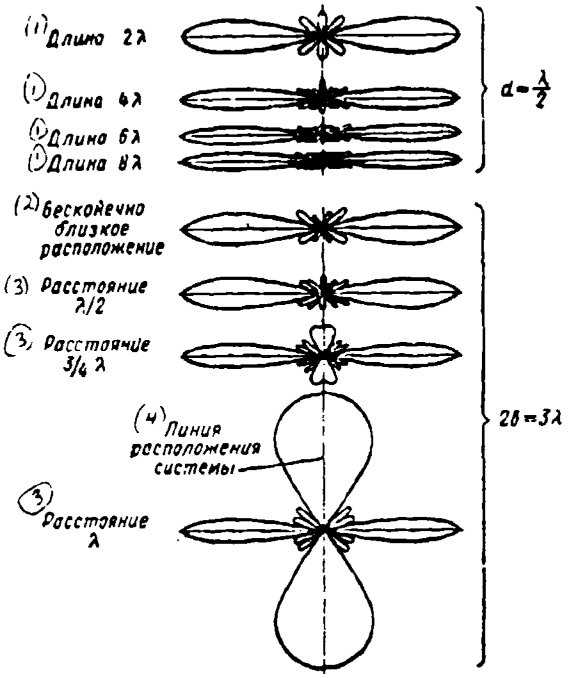

Fig. 3.36. Effect of the number of the vertical wire antennas and distance between them on directional characteristic: $d = \lambda/2$ - the effect of length 2 v on radiation pattern with the distance between the vertical wire antennas $\lambda/2$; $2b = 3\lambda$ - the effect of the distance between the vertical wire antennas on radiation pattern at the overall length of system, equal to 3λ.

Key: (1). Length. (2). Infinitely close location. (3). Distance. (4). Line of the location of system.

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In this case, it is accepted that $F_1(\theta)=1$, i.e. that the separate antennas do not possess the directional reception. On Fig. 3.37, are given for this system the directive gains D depending on 2b/λ. On the basis of curves (Fig. 3.36) it is possible to make following conclusions. The fundamental maximum of diagram is always directed along perpendicular to the line of antenna location. Diagrams are symmetrical with respect to the line of antenna location and with respect to perpendicular to it.

Simultaneously with the main lobe of radiation are minor lobes. Their number with change θ from 0 to 90° equal to number whole waves in complete separation 2b. The level of the maximums of minor lobes is designed from the formula

$$F_n(\theta)=\frac{1}{n\sin\left(\frac{2k+1}{n}\frac{\pi}{2}\right)}. \qquad (3.76)$$

With large n formula (3.76) is simplified:

$$F_n(\theta)=\frac{2}{(2k+1)\pi}$$

and for the maximums of minor lobes we obtain consecutively values of

0.212, 0.128, of 0.091 so forth from the level of major lobe.

With an increase in the separation between extreme antennas, the width of the main lobe of radiation decreases. The width of major lobe with the drop of amplitude to zero for large n is equal to

$$2\gamma_0 \approx \frac{2(n-1)}{n}\frac{\lambda}{2b} \approx 114,6\frac{\lambda}{nd}, \text{ deg.} \qquad (3.77)$$

The width of major lobe with the drop of power two times with the same condition will be $2\gamma_{0.5} \approx 50,8\frac{\lambda}{nd}$, deg.

With that which was assigned 2b width of major lobe weakly depends on the number of antennas, decreasing with increase in d. However, considerably cannot be increased d, since with an increase of d, i.e., with decrease in n with that which was assigned 2b, increases relative value of minor lobes. With $d \geqslant \lambda$ minor lobes are equal to the main thing, i.e., appear supplementary principal maximums.

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Sometimes is placed the problem of suppression to the determined side-lobe level of radiation pattern. By characteristic the optimum antenna of system, which possesses the minimum width of major lobe with the required side-lobe level and equidistance between antennas, is Chebyshev polynomial degree n - 1. The voltages, removed from antennas, in this case decrease to the edges of antenna.

This they achieve, for example, by introduction into the antennas of the voltage dividers; then the effective height/altitude of system decreases.

The suppression of minor lobes it is possible to also achieve (with the identical voltages, removed from antennas) by the establishment of the different distances between antennas, which increase to the edges of system.

The degree of the suppression of minor lobes in principle can be any.

The calculations of the voltages, removed from antennas, in the first case also of the required distances between antennas in the second case are given in the courses of antennas [3.2, 3.3].

During a decrease in the minor lobes, is expanded the main lobe of radiation and decreases directive gain.

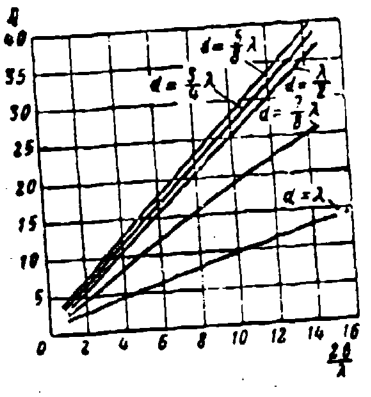

Fig. 3.37. Directive gains of system from n of antennas with respect to single emitter.

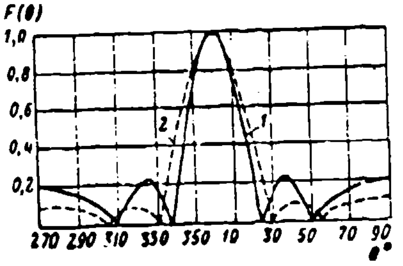

Fig. 3.38. Directional characteristic: 1 - uniform grating; 2 - grating with Chebyshev distribution.

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On Fig. 3.38 for a comparison, are given the radiation patterns of uniform grating and grating with the suppression of minor lobes to

0.1 with n = 5 and d = 0.5λ.

It is possible n of antennas in question to break into two cophasal groups by n/2 antennas and to connect both groups so that the voltages of groups were subtracted. Then we obtain radiation pattern with zero along perpendicular to the line of antennas. 11 directional characteristic will be

where F_1 (θ, β) is directional characteristic of single antenna;

F_2 (θ, β) - the standardized/normalized characteristic n/2 of the broadside antenna arrays;

F_3 (θ, β) - the standardized/normalized characteristic of two groups of the antennas, connected on differential principle.

$$F_2(\theta,\ \beta)=\frac{\sin\left(\frac{n}{4}\,m\,d\sin\theta\cos\beta\right)}{\frac{n}{2}\sin\left(\frac{1}{2}\,m\,d\sin\theta\cos\beta\right)},\qquad(3.79)$$

$$F_3(\theta,\ \beta)=\sin\left(\frac{md_2}{2}\sin\theta\cos\beta\right).\qquad(3.80)$$

Here D_2 = d n/2 is a distance between the middles of both groups of antennas.

After substituting formula (3.79), (3.80) in (3.78), we will obtain

$$F_p(0,\ \beta)=F_1(0,\ \beta)\frac{1-\cos\left(\frac{n}{2}md\sin\theta\cos\beta\right)}{n\sin\left(\frac{1}{2}md\sin\theta\cos\beta\right)}. \qquad (3.81)$$

Differential directional characteristic (3.81) is symmetrical relative to the line of antennas and standard to the line of antennas; it has minor lobes number and maximums of which depend on separation d and the number of antennas of system.

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The main lobe of radiation seemingly bifurcated itself - instead of the maximum, directed along perpendicular to the line of antennas, it has zero value in the direction of this perpendicular.

The important characteristic of differential radiation pattern, which determines the accuracy of direction finding, is mutual conductance in zero $F'_p(0)$. In by usual antenna to system with equidistances between antennas with the identical voltages, removed from separate antennas, for the number of antennas

$$n>5 \quad \text{and} \quad \frac{d}{\lambda}<0{,}5\ F'_p(0)=0{,}7\frac{2\pi b}{\lambda}.$$

For an increase in mutual conductance $F'_{p}(0)$ it is necessary to remove/take with the antenna of system the different voltages, which decrease to the edges of its halves. Directional characteristic of the optimum system, which has the maximum value of slope/transconductance in zero on the assigned side-lobe level and with the minimum width of major lobe of diagram, is the polynomial of Akhiyezer of degree n - 1.

Work [3.10] gives calculation by this antenna of system and are investigated its parameters. It is shown, that the maximally attainable slope/transconductance, when $d/\lambda \leqslant 1/2$, will be

$$F'_{p}(0)_{\text{макс}} = \frac{1}{\sin\frac{\pi d}{\lambda}} \frac{2\pi b}{\lambda}.$$

With that which was assigned 2b slope/transconductance $F'_{p}(0)$ increases with increase in n.

With $d/\lambda > 1/2$ slope/transconductance $F'_{p}(0)_{\text{макс}} = \frac{2\pi b}{\lambda}$, i.e. has the same value, as of system of two antennas.

A deficiency/lack in the optimum systems is their narrow band coverage.

A system of the vertical wire antennas can be utilized as rotary for determining the direction of radio stations in the minimum or in the maximum of radiation pattern.

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In order to obtain the directional reception only on one hand, one should to use reflector, i.e., establish/install a series of wires, parallel to antennas, in that side from by the antenna of system, whence is undesirable reception. Distance of reflector usually takes equal to 1/4 average wavelengths by the antenna of system. The wires of reflector must on height/altitude project/emerge beyond the limits of antennas at least to value d_p - the distance between antennas and the reflector. To the same value the width of reflector to each side must exceed the separation between extreme antennas.

The effect of reflector, which is located at a distance d_p from a series of the vertical wire antennas, is considered by the facts that in calculation instead of each antenna of system takes two antennas at a distance $2d_n$ one from another and with phase displacement 180° at e. d. the s. of antennas. In order to obtain directional characteristic of this system in front of reflector, it follows directional characteristic of the system of antenna without reflector (3.74) or (3.81) to multiply by sin $(md_p \cos 0)$.

If we between the separate vertical wire antennas of a series (Fig. 3.35) introduce phase displacement ψ, then radiation pattern instead of (3.74) and (3.81) they will be expressed

$$F(\theta,\ \beta,\ \psi)=F_1(\theta,\ \beta)\frac{\sin\left[\frac{n}{2}(m\,d\sin\theta\cos\beta-\psi)\right]}{n\sin\left[\frac{1}{2}(m\,d\sin\theta\cos\beta-\psi)\right]}, \quad (3.82)$$

$$F_p(\theta,\ \beta,\ \psi)=F_1(\theta,\ \beta)\frac{1-\cos\left[\frac{n}{2}(m\,d\sin\theta\cos\beta-\psi)\right]}{n\sin\left[\frac{1}{2}(m\,d\sin\theta\cos\beta-\psi)\right]}. \quad (3.83)$$

Equalizing zero argument of the numerator of expressions (3.82) and (3.83), we obtain, that the maximum or the minimum of the main lobe of radiation with phase displacement ψ_m is turned to angle θ_m, determined from the condition

$$\psi_m=m\,d\sin\theta_m\cos\beta. \quad (3.84)$$

The angle between the opposite maximums or the minimums of diagram in this case is equal to 180°$\pm 2\theta_m$, i.e. it appears the fracture of the centerline of radiation pattern relative to the line of antennas.

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Furthermore, diagrams become asymmetric relative to centerline.

Changing phase displacement ψ_m, it is possible, utilizing a motionless series of the vertical wire antennas, to rotate diagram of the directivity of system and according to maximum or minimum of diagram to determine direction in radio station. Deficiency/lacks in the system are fracture of the centerline of radiation pattern during its rotation and asymmetry of diagram. The first deficiency/lack drops off during the application/use of a reflector, since is eliminated one of the maximums of reception. Asymmetry of diagram limits φ_m.

For obtaining the rotation of radiation pattern, phase displacement ψ_m should design from (3.84) for certain mean angle of the slope of a front of wave β_c and calibrate phase displacement in the degrees of the rotation of the maximum (or the minimum) of radiation pattern.

With a change in the angle of the slope of a front of wave β, is obtained the high-altitude error in the determination of the maximum (or the minimum) of radiation pattern.

Actually, at the angle β, different from β_c, for which are designed phase displacements, the measured angle $\varphi_и$ will not be equal real $\varphi_д$, since

$$\sin\varphi_д\cos\beta=\sin\varphi_и\cos\beta_c. \qquad (3.85)$$

Will be obtained the error in the determination of direction $\Delta=\varphi_и-\varphi_д$.

which it is designed with the aid of formula (3.35).

On Fig. 3.39, are given to the dependence $\varphi_{\text{н}}$ on $\varphi_{\text{л}}$ with different β from 0 to 60° and $\beta_{c}=0^{\circ}$.

If phase displacements of system are designed for any other angle of the slope of a front of wave $\beta_{c}\neq0^{\circ}$, then high-altitude error for an angle of incidence β is equal to a difference in the ordinates of curves, that correspond to the values of angles of incidence β and β_{c} at the assigned values $\varphi_{\text{л}}$, or a difference in the abscissas of the same curves at the assigned values $\varphi_{\text{н}}$.

During calculation by the described antenna of system, it is necessary to consider mutual antenna resistance, which is determined from the method of those induced by emf [3.1, 3.4]. If antenna system is designed from n of antennas, then they are usually establish/installed by n + 2 antennas.

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Extreme antennas are not included in diagram and serve only for the creation of the identity of the resistor/resistances of all antennas.

System with acute/sharp radiation pattern as any other antenna system, is characterized by the radiation resistance, efficiency and KND [directivity factor]. These parameters determine the sensitivity of radio direction finder.

The methods of the calculation of the indicated parameters by the antenna of system are set forth with the courses of antennas.

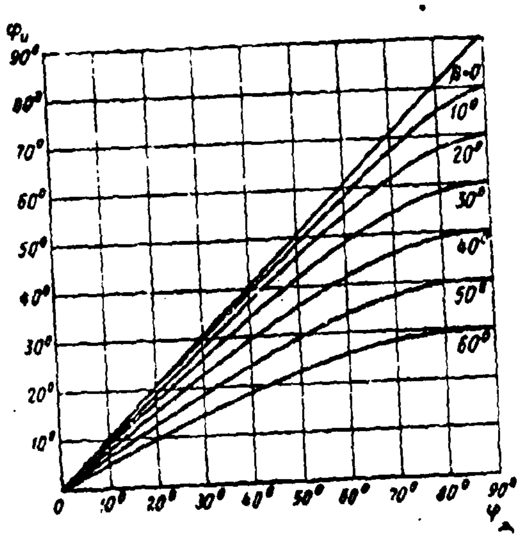

Fig. 3.39. The dependence $\varphi_{и}$ on $\varphi_{л}$ with different β (it is acceptel that $\beta_c = 0$).

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Circular antenna systems with acute/sharp directional characteristic.

For the realization of smooth rotation to 360° acute/sharp directional characteristic in motionless antennas, is applied antenna system from arranged on circumference n of the vertical wire antennas. A circular antenna system can be utilized with reflector and without reflector.

The principle of formation/education and rotation of radiation pattern in by the circular antenna to system includes the following (Fig. 3.40). From n of the antennas of system, are selected m of the antennas which form direction-finding group. In the antennas of group with the aid of antenna commutator, is introduced to the circuit of a time delay in this value (11'/c, 22'/c, 33'/c ...., where c is velocity of propagation of radio waves) in order to make even the phases of emf of the antennas of group for direction OO' and seemingly to lead antennas to those who were arrange/located in straight line AA_1. By the central line OO' of the antenna of group they are divided into two subgroups. Emf of the antennas of right and the left of subgroups store/add up into two voltages E_1 and E_1 which for direction finding they are summarized, or from E_1 it is deducted E_2.

During the rotation of antenna commutator, change the delays in antennas and antennas themselves so are changed over, that the center of line AA_1 seemingly rotated synchronously with the rotation of commutator in internal broken circumference (Fig. 3.40), the line of antennas occupying positions BB_1, CC_1 so forth.

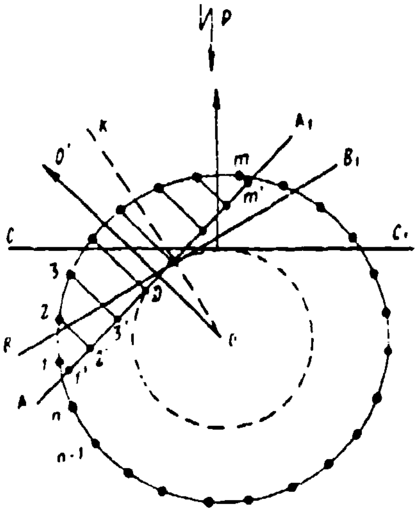

Fig. 3.40. Principle of the use of a circular system of antennas.

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When the line of antennas CC_1 is perpendicular to direction in of radio station P, is obtained maximum voltage according to total directional characteristic and zero voltage according to differential characteristic.

To Fig. 3.41, are shown the total and differential radiation patterns of direction-finding group. Minor lobes in figure for simplification are not display. Antenna commutator consists of motionless stator and the rotatable rotor (Fig. 3.42).

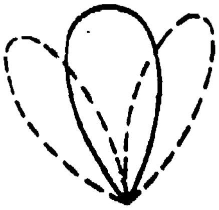

Fig 3.41

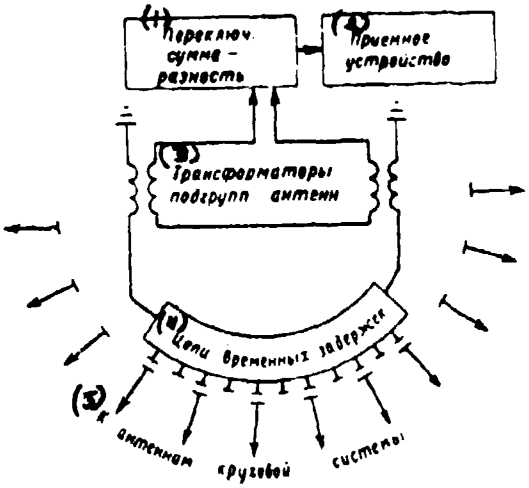
(1) Переключ. сумма - разность
(2) Приемное устройство
(3) Трансформаторы подгрупп антенн
(4) Цепи временных задержек
(5) к антеннам круговой системы

Fig 3.42

Fig. 3.41. Total and differential radiation patterns: ——— total diagram, - - - differential diagram.

Fig. 3.42. Diagram of antenna commutator for by the circular antenna of system.

Key: (1). Switch is sum of difference. (2). Receptor. (3). Transformers of the subgroups of antennas. (4). Circuits of time delays. (5). To the antennas of circular system.

[Page 143] Stator-rotor unit have platas, serving for the creation of capacitive coupling between them.

To the plates of stator, are connected the antennas, to the plates of the rotor, which has two symmetrical halves, serving for the creation of two halves of the group of antennas, are wired circuits of time delays.

The latter can be fulfilled in the form of the cut cables or artificial line. Delay units can be utilized separate for each plate of rotor or can be applied common/general/total for each half of rotor. The electrical length of chain of time delays is designed on the basis of geometry by the antenna of system for certain initial angle of the slope of a front of wave β_r. During the application/use of common/general/total delay circuits, the load from antennas must be 4-5 times the more than wave impedance of circuit. Then the disturbance/breakdown of the mode of the traveling wave small is retained the electrical length of line.

The number of plates of stator is equal to the number of antennas (n). The number of plates of rotor, which are necessary to one plate of stator, or the multiplicity of the plates of rotor relative to the plates of stator is selected based on the instrument errors of antenna commutator (see §8.9). Total number of plates of rotor depends on the number of antennas in direction-finding group. Separately store/add up the voltages of right and by the left of subgroups E_1 and E_2. By a special switch is realized addition or the subtraction of voltages E_1 and E_2.

Direction finding consists in the searching of maximum or minimum of reception. Angle on the scale of antenna of accumulator corresponds to azimuth.

Direction finding can also be realized without the rotation of antenna commutator according to the sum-and-difference method whose principle is described into §2.3. For this, antenna commutator is establish/installed so as to lead the direction-finding group of antennas, for example, to straight line AA_1 (see Fig. 3.40). The central line OO_1 in this case is the zero reference line of bearing. The sum of the voltages of two subgroups of antennas AD and $DA_1 (U_\Sigma)$ is fed to one pair of the plates of cathode-ray tube, for example to longitudinal plates.

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Voltage difference of these subgroups of antennas (U_Δ). are phase shifted by 90°, is fed to the second pair of the plates of tube, respectively to horizontal plates. Direction finding is realized in certain sector $2\theta_{\text{макс}}$.

Let us replace each subgroup of antennas with one equivalent antenna with effective height $h_д$.

Let us designate:

$2b_э$ - the equivalent separation of two antennas, which replace the subgroups of antennas;

θ' - the azimuth of radio station with line OO_1.

Then in accordance with that which was presented is earlier the amplitude of the sum of the voltages of the antennas

$$U_\Sigma = 2Eh_д \cos(mb_э \cos\beta \sin\theta'),$$

the amplitude of voltage difference of the antennas

$$U_\Delta = 2Eh_д \sin(mb_э \cos\beta \sin\theta').$$

On cathode-ray tube is counted off the angle α, determined by expression $\mathrm{tg}\,\alpha = \frac{U_\Delta}{U_\Sigma} = \mathrm{tg}(mb_э \cos\beta \sin\theta')$, or $\alpha = mb_э \cos\beta \sin\theta' = 0{,}5(2mb_э \cos\beta \sin\theta')$.

Angle on cathode-ray tube is equal to the half of a difference is phase. Emf. in the antennas ψ, since $\psi = 2mb_0 \cos\beta \sin\theta'$.

The scaling factor $\kappa_п$ from angle on cathode-ray tube α to azimuth θ' will be (at low values θ', when sin θ' ≈ θ').

$$\kappa_п = mb_э \cos\beta.$$

When β = 0 $\kappa_п = \frac{2\pi b_э}{\lambda}$ and $\theta' = \frac{\alpha}{\kappa_п}$.

The sector of direction finding (2.18) is determined by the expression

$$\theta'_{\text{макс}} < \frac{\frac{\pi}{2}}{K_{\text{п}_1}} \text{ or } 2\theta'_{\text{макс}} < \frac{28.6\lambda}{b_{\text{э}}}, \text{ deg.}$$

To performance calculation of the directivity of circular antenna systems is dedicated a series [3.12, 3.14, 3.15].

Let us determine for an example of the expression of directional characteristic for by the antenna of system without reflector in the case when for the formation of directional characteristic simultaneously are utilized all antennas [3.12].

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We count off azimuth θ to radio station from diameter by the antenna of system, of the corresponding to the position of the line of symmetry antenna commutator (CC_1 on Fig. 3.43). Let us designate by α_0 the angle between direction of CC_1 and by the radius, carried out to the nearest to it clockwise antenna which let us consider the n-th, moreover α_0 it varies from 0 to 2π/n.

If E_0 is strength of field at the center of the system: b - the radius of a circle of the arrangement/permutation of antennas, then emf induced in the m antenna, will be

$$\dot{E}_m = E_0 h_{\partial 0} \exp\left[j \frac{2\pi}{\lambda} b \cos\beta \cos\left(\theta - \frac{2\pi}{n} m - \alpha_0 \right) \right].$$

In the case in question line AA_1, to which are led all antennae by means of the introduction of the phase delays in emf of antennas, is tangential to the circumference of the arrangement/permutation of antennas (see Fig. 3.43). In emf of the m antenna, in order to lead the phase of emf of this antenna to the phase, which corresponds to line AA_1, is introduced phase displacement ψ_m:

$$\psi_m = \frac{2\pi}{\lambda} b \cos\beta_0 \left[1 + \cos\left(\frac{2\pi}{n} m + \alpha_0 \right) \right].$$

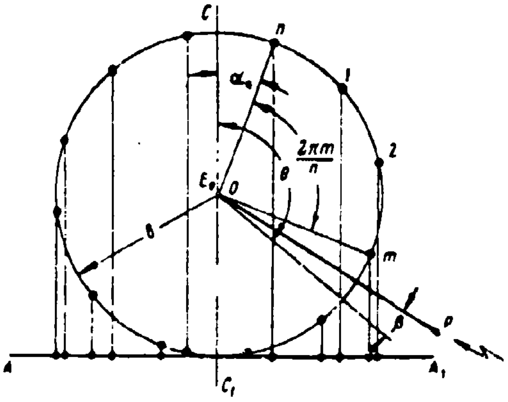

Fig. 3.43. Bringing by the antenna of circular system to linear.

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The voltage, removed from the m antenna, taking into account φ_m is equal

$$\dot{U}_m = E_0 h_{\pi o} \exp j \left\{ \frac{2\pi}{\lambda} b \left[\cos\beta \cos\left(\theta - \frac{2\pi}{n} m - \alpha_0\right) - \right.\right.$$
$$\left.\left. - \cos\beta_0 \cos\left(\frac{2\pi}{n} m + \alpha_0\right)\right] - \frac{2\pi}{\lambda} b \cos\beta_0 \right\}. \qquad (3.86)$$

The last/latter term of exponent does not depend on α_0 and θ. We convert the depending on α_0 and θ part of the exponent:

$$\cos\beta\cos\left(\theta-\frac{2\pi}{n}m-\alpha_0\right)-\cos\beta_0\cos\left(\frac{2\pi}{n}m+\alpha_0\right)=$$
$$=\cos\left(\frac{2\pi}{n}m+\alpha_0\right)(\cos\beta\cos\theta-\cos\beta_0)+$$
$$+\sin\left(\frac{2\pi}{n}m+\alpha_0\right)\cos\beta\sin\theta=A\cos\left(\gamma-\frac{2\pi}{n}m-\alpha_0\right),$$

where $$A=\sqrt{\cos^2\beta+\cos^2\beta_0-2\cos\beta\cos\beta_0\cos\theta} \quad (3.87)$$

and $$\operatorname{tg}\gamma=\frac{\cos\beta\sin\theta}{\cos\beta\cos\theta-\cos\beta_0}. \quad (3.88)$$

After substituting (3.87) and (3.88) in (3.86), we will obtain

$$\dot{U}_m=E_0h_{e0}\exp\left\{j\left[\frac{2\pi}{\lambda}bA\cos\left(\gamma-\frac{2\pi}{n}m-\alpha_0\right)\right]-\right.$$
$$\left.-j\frac{2\pi}{\lambda}b\cos\beta_0\right\}. \quad (3.89)$$

It is decomposed (by 3.89) in Fourier series - Bessel analogous with expansion (3.49), moreover factor $e^{-j\frac{2\pi}{\lambda}b\cos\beta_0}$ is not considered, since it does not depend on θ and α_0:

$$\dot{U}'_m=E_0h_{e0}\left\{J_0\left(\frac{2\pi}{\lambda}bA\right)+\right.$$
$$\left.+2\sum_{p=1}^{\infty}j^pJ_p\left(\frac{2\pi}{\lambda}bA\right)\cos\left[p\left(\gamma-\frac{2\pi}{n}m-\alpha_0\right)\right]\right\}. \quad (3.89')$$

Let us determine with n even the total voltage, removed from all antennas:

$$\dot{U}_\Sigma=\sum_{m=1}^{n}\dot{U}'_m=E_0h_{e0}\left\{nJ_0\left(\frac{2\pi}{\lambda}bA\right)+\right.$$

$$+2\sum_{m=1}^{n}\sum_{p=1}^{\infty} j^{p}J_{p}\left(\frac{2\pi}{\lambda}bA\right)\cos\left[p\left(\gamma-\frac{2\pi}{n}m-\alpha_{0}\right)\right]\Big\}$$

or

$$U_{\Sigma}=E_{0}h_{an}n\Big\{J_{0}\left(\frac{2\pi}{\lambda}bA\right)+$$
$$+2\sum_{v=1}^{\infty}(-1)^{\frac{vn}{2}}J_{vn}\left(\frac{2\pi}{\lambda}bA\right)\cos\left[vn(\gamma-\alpha_{0})\right]\Big\}. \quad (3.90)$$

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With addition it is accepted into consideration, which

$$\sum_{m=1}^{n}\cos\left[p\left(\gamma-\frac{2\pi}{n}m-\alpha_{0}\right)\right]=n\cos\left[vn(\gamma-\alpha_{0})\right]$$

with p = vn is equal to zero at other values of p.

Expression (3.90) determines total directional characteristic by the antenna of system. Usually it is possible to be restricted two first terms of the series

$$U_{\Sigma}=E_{0}h_{an}n\Big\{J_{0}\left(\frac{2\pi}{\lambda}bA\right)+2(-1)^{\frac{n}{2}}J_{n}\left(\frac{2\pi}{\lambda}bA\right)\cos\left[n(\gamma-\alpha_{0})\right]\Big\}. \quad (3.91)$$

So that it would be possible to disregard the second term of expression (3.91), it is necessary to select the number of antennas, so that the distance between them will be

$$d < (0,45 \div 0,5)\lambda. \qquad (3.92)$$

This condition determines the necessary number of antennas with that which was assigned $\frac{b}{\lambda_{\text{мин}}}$ and, on the contrary, causes the permissible sense $\frac{b}{\lambda_{\text{мин}}}$ with the assigned number of antennas. Then total directional characteristic is determined by the expression

$$F_{\Sigma}(\theta, \beta) = J_0\left(\frac{2\pi}{\lambda} bA\right). \qquad (3.93)$$

In Table 3.1 are designed during fulfilling of requirement (3.92) for value $\frac{b}{\lambda_{\text{мин}}}$ for the different number of antennas n when $\beta_c = 20°$. In them are made angles Φ_0, which limit the main lobe of radiation.

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When $\beta = \beta_c$, that $A = 2\cos\beta_c \sin\frac{\theta}{2}$ and $\gamma = \pi/2 + \theta/2$.

In this case the expression for total directional characteristic takes the form

$$F_{\Sigma}(\theta, \beta = \beta_c) = J_0\left(\frac{2\pi}{\lambda} 2b\cos\beta_c \sin\frac{\theta}{2}\right). \qquad (3.94)$$

The amplitudes of the minor lobes of characteristic (3.94) have relative values 0.403; 0.3; 0.25; 0.21 so forth. On Fig. 3.44, are constructed for an example total directional characteristic ~~when~~ $\beta=\beta_c=20°$ for different ratios 2b/λ and the different number of antennas N [3.12].

Table 3.2 gives the values angles limiting the minor lobes of directional characteristic in plane $\beta=\beta_c=0°$.

With θ = 0 we obtain expression for total directional characteristic in the vertical plane

$$F(\theta=0,\beta)=J_0\left[\frac{2\pi}{\lambda}b(\cos\beta-\cos\beta_c)\right]. \qquad (3.95)$$

On Fig. 3.45, is constructed for an example directional characteristic in vertical plane for b/λ = 0.75 and $\beta_c=20°$.

Table 3.1. Parameters by the circular antenna of system depending on the number of antennas n when $\beta_c = 20°$

(1) Число антенн n	6	8	12	16	20	24	28	32	36	40
$\frac{2b}{\lambda_{мин}} \times \cos\beta_c$	0,955	1.27	1,91	2,54	3,18	3,8	4,45	5,1	5,75	6,37
(2) Граничный угол главного лепестка суммарной характеристики, *град*	50,4	38	25	18	15	12.3	10,5	9,15	8,1	7,3

Key: (1). Number of antennas n. (2). Limiting angle of major lobe of total characteristic, deg.

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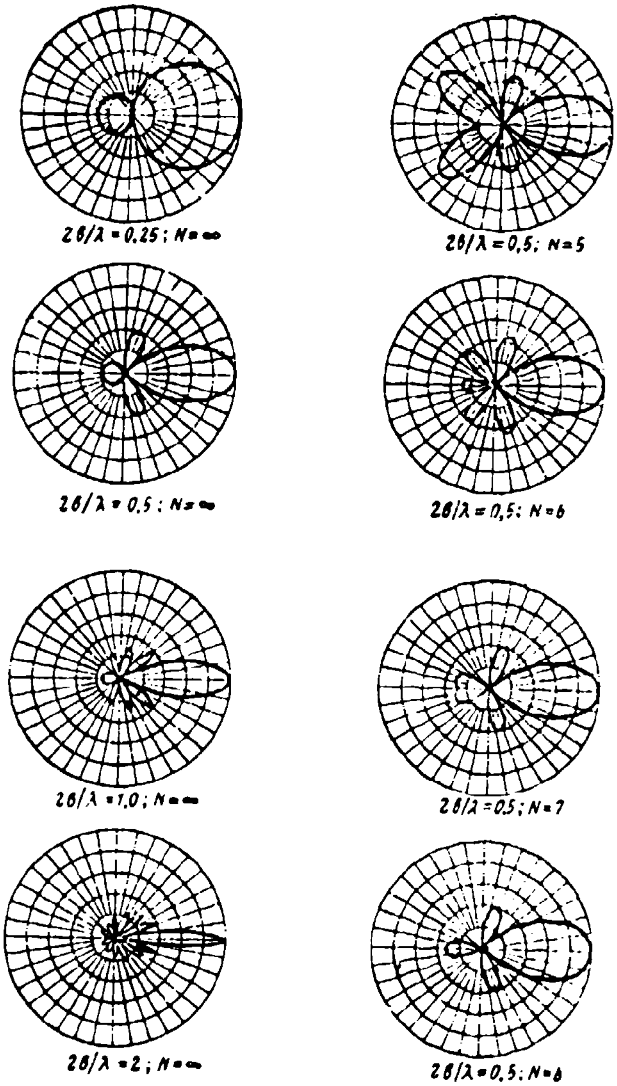

$2б/\lambda = 0{,}25;\ N = \infty$

$2б/\lambda = 0{,}5;\ N = 5$

$2б/\lambda = 0{,}5;\ N = \infty$

$2б/\lambda = 0{,}5;\ N = 6$

$2б/\lambda = 1{,}0;\ N = \infty$

$2б/\lambda = 0{,}5;\ N = 7$

$2б/\lambda = 2;\ N = \infty$

$2б/\lambda = 0{,}5;\ N = 6$

Fig. 3.44. Total directional characteristic of the circular system of antennas.

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In Table 3-3 are designed the most interesting values of directional characteristic in the vertical plane when $\beta_c = 20°$, namely of characteristic value with $\beta = 0$, angles $\beta'_{макс}$ corresponding to the weakening of major lobe to 0.7 and $\beta''_{макс}$ on the boundary of major lobe (with zero values of characteristic).

Table 3.2.

Values of the angles, which correspond to the maximums also of zero lug/lobes of total directional characteristic by the circular antenna of system in plane $\varphi=\beta_0=0^\circ$

(1) Наименование лепестка	(2) Амплитуда лепестка	(3) Отношение $\frac{b}{\lambda}\cos\beta_c$					
		0,25	0,5	1,0	1,5	2,0	2,5
(4) Главный лепесток характеристики направленности	1 0	0° 100°	0° 45°	0° 22°	0° 15°40′	0° 11°	0° 8°40′
(5) Первый боковой лепесток	0,403 0		75°20′ 123°20′	35°40′ 52°20′	23°40′ 34°10′	17°40′ 25°30′	14° 20°20′
(6) Второй боковой лепесток	0,3 0			69° 87°30′	44°20′ 55°	32°50′ 39°20′	26° 32°
(7) Третий боковой лепесток	0,25 0			108° 140°40′	65°20′ 77°40′	48° 56°	38° 44°
(8) Четвертый боковой лепесток	0,22 0			180° (0,16)	90° 105°	64° 73°	50°20′ 57°
(9) Пятый боковой лепесток	0,2 0				122° 144°	82° 93°	63°20′ 70°30′

Key: (1). Designation of lobe. (2). Amplitude of lug/lobe. (3). Relation b/λ cos β_c. (4). Major lobe of directional characteristic. (5). First minor lobe. (6). Second minor lobe. (7). Third minor lobe. (8). Fourth minor lobe. (9). Fifth minor lobe.

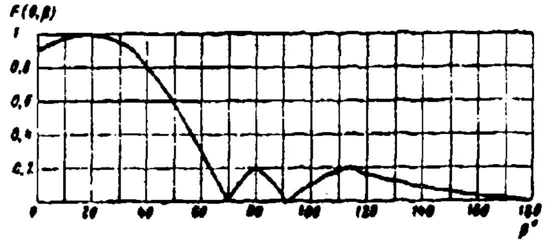

Fig. 3.45. Directional characteristic in vertical plane.

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Let us determine the separation of two antennas $2b_э$, equivalent by the circular antenna to system in the relation to the width of major lobe of total directional characteristic.

We set/assume $\beta=\beta_c=0°$. Conditions for obtaining the boundary of major lobe of total directional characteristic of circular system and system of two antennas will be respectively

$$\frac{4\pi}{\lambda} b \sin\frac{\theta_0}{2} = 2{,}43 \quad \text{and} \quad \frac{2\pi}{\lambda} b_э \sin\theta_0 = \frac{\pi}{2}.$$

We take, that $\sin\theta_0 \approx \theta_0, \ \sin\frac{\theta_0}{2} \approx \frac{\theta_0}{2}$.

Then $b_э = \frac{1{,}57}{2{,}43} b = 0{,}645\, b$

or $2b_a = 1{,}29b$, where 2b is a separation (diameter) of antennas.

Let us find expression for differential directional characteristic. Voltage difference, removed from diametrically opposite antennas the m-th and (m + n/2)-1 of (3.89'), will be

$$U_{m\Delta} = U_m - U_{m+\frac{n}{2}} =$$

$$= 2E_0 h_{eo} \sum_{q=1}^{\infty} j^q J_q\left(\frac{2\pi}{\lambda} bA\right)\left\{\cos\left[q\left(\gamma - \frac{2\pi}{n} m - \alpha_0\right)\right] - \right.$$

$$\left. - \cos\left[q\left(\gamma - \frac{2\pi}{n} m - \pi - \alpha_0\right)\right]\right\} =$$

$$= -4E_0 h_{eo} \sum_{q=1}^{\infty} j^q J_q\left(\frac{2\pi}{\lambda} bA\right) \times$$

$$\times \sin\left[q\left(\gamma - \frac{2\pi}{n} m - \frac{\pi}{2} - \alpha_0\right)\right] \sin\left(q\frac{\pi}{2}\right). \qquad (3.96)$$

Table 3.3.

Characteristic values of total directional characteristic by the circular antenna of system in vertical plane $F_{\Sigma}(0,\beta)$ when $\beta_0 = 20°$.

$\frac{b}{\lambda}$	0,5	1	1,5	2	2,5	3,0
$F_{\Sigma}(0,\beta_0)$	0,99	0,97	0,92	0,86	0,79	0,71
$\beta'_{\text{макс}}$	56°	41°	36°	32°	30°	29°
$\beta''_{\text{макс}}$	80°	57°	47°	41°	38°	36°

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Terms (3.96) differ from zero only at the odd values of q, since at the even values of q = 2p $\sin(2p\frac{\pi}{2}) = 0$. Therefore expression (3.96) can be rewritten, set/assuming q = 2p - 1, in the form

$$U_{m\Delta} = -4E_0 h_{c0}\sum_{p=1}^{\infty} j^{2p-1} J_{2p-1}\left(\frac{2\pi}{\lambda} bA\right)\times$$

$$\times\sin\left[(2p-1)\left(\gamma-\frac{2\pi}{n}m-\alpha_0\right)-p\pi+\frac{\pi}{2}\right]\sin\left[(2p-1)\frac{\pi}{2}\right]=$$

$$= -j4E_0 h_{e0}\sum_{p=1}^{\infty}(-1)^{p-1} J_{2p-1}\left(\frac{2\pi}{\lambda} bA\right)\times$$

$$\times\cos\left[(2p-1)\left(\gamma-\frac{2\pi}{n}m-\alpha_0\right)-p\pi\right]\sin\left[(2p-1)\frac{\pi}{2}\right]$$

or

$$U_{m\Delta}=j4E_0h_{e0}\sum_{p=1}^{\infty}(-1)^{p-1}J_{2p-1}\left(\frac{2\pi}{\lambda}bA\right)\times$$
$$\times\cos\left[(2p-1)\left(\gamma-\frac{2\pi}{n}m-\alpha_0\right)\right].$$

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Voltage difference from all antennas U_Δ will be (see Fig. 3.43)

$$U_\Delta=\sum_{m=0}^{\frac{n}{2}-1}U_{m\Delta}=j4E_0h_{e0}\sum_{p=1}^{\infty}(-1)^{p-1}J_{2p-1}\left(\frac{2\pi}{\lambda}bA\right)\times$$
$$\times\sum_{m=0}^{\frac{n}{2}-1}\cos\left[(2p-1)\left(\gamma-\frac{2\pi}{n}m-\alpha_0\right)\right].$$

where

$$\sum_{m=0}^{\frac{n}{2}-1}\cos\left[(2p-1)\left(\gamma-\frac{2\pi}{n}m-\alpha_0\right)\right]=$$
$$=\cos[(2p-1)(\gamma-\alpha_0)]\sum_{m=0}^{\frac{n}{2}-1}\cos\left[(2p-1)\frac{2\pi}{n}m\right]+$$
$$+\sin[(2p-1)(\gamma-\alpha_0)]\sum_{m=0}^{\frac{n}{2}-1}\sin\left[(2p-1)\frac{2\pi}{n}m\right].$$

After designating

$$x=(2p-1)\frac{2\pi}{n},$$

after transformations we will obtain (see (1.342) [1.13])

$$\sum_{m=0}^{\frac{n}{2}-1}\cos\left[(2p-1)\left(\gamma-\frac{2\pi}{n}m-\alpha_0\right)\right]=$$

$$=\cos[(2p-1)(\gamma-\alpha_0)]+\sin[(2p-1)(\gamma-\alpha_0)]\times$$

$$\times\cos\left[(2p-1)\frac{\pi}{n}\right]\operatorname{cosec}\left[(2p-1)\frac{\pi}{n}\right]=$$

$$=\sin\left[(2p-1)\left(\gamma-\alpha_0+\frac{\pi}{n}\right)\right]\operatorname{cosec}\left[(2p-1)\frac{\pi}{n}\right].$$

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It summed up voltage $U_{m\Delta}$ from all antennas, we will obtain

$$U_{\Delta}=j4E_0h_{e0}\sum_{p=1}^{\infty}(-1)^{p-1}J_{2p-1}\left(\frac{2\pi}{\lambda}b\dot{A}\right)\times$$
$$\times\sin\left[(2p-1)\left(\gamma+\frac{\pi}{n}-\alpha_0\right)\right]\operatorname{cosec}\left[(2p-1)\frac{\pi}{n}\right]. \quad (3.97)$$

The standarilized/normalized differential characteristic has the expression

$$F_{\Delta}(0,\beta)=\frac{4}{n}\sum_{p=1}^{\infty}(-1)^{p-1}J_{2p-1}\left(\frac{2\pi}{\lambda}b\dot{A}\right)\times$$
$$\times\sin\left[(2p-1)\left(\gamma+\frac{\pi}{n}-\alpha_0\right)\right]\operatorname{cosec}\left[(2p-1)\frac{\pi}{n}\right]. \quad (3.98)$$

An example of differential radiation pattern is given in Fig. 3.46.

The presence of reflector is led to the peaking of directional characteristic. Reflector is establish/installed from antenna at a distance, equal approximately of 1/4 average wavelengths of operating range.

The method of performance calculation of the directivity of the group of the antennas of circular system with reflector is presented in [3.15].

Figures 3.47 gives for a comparison major lobes of total and differential directional characteristic for by the circular antenna of system with reflector on wave 18.8 m and cosinusoidal characteristic.

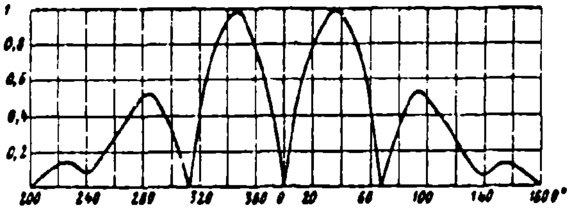

Fig. 3.46. Difference directional characteristic of circular system of antennas.

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Parameters by the circular antenna by the antenna of system (German radio direction finder "Vullenveber" of development 1940-1945): n = 40, m = 8, 2b = 120 m, reflector it is establish/installed at a distance 12.5 m of antennas; antenna vertical with a diameter of 3.4 m, by the height/altitude of 7.5 m, with capacitive load above; frequency band 6-15 MHz [3.14].

3.12. Antennas with logarithmic structure.

So that the antenna will be wide-range, i.e., it has the not changing with change frequencies entry impedance and radiation pattern, it must have the smoothly and equally changing along the length section and the equivalent length, inversely proportional to frequency. This requirement satisfy the infinitely extended antennas whose form is determined only by smoothly changing angular dimensions. The problem of developing the wide-range antenna of finite length is finding such structure whose form is determined by angular dimensions and the behavior of final part of which at frequencies higher than certain boundary approaches behavior of infinite structure.

From the flat antennas of such property possesses the logarithmic spiral (Fig. 3.48) whose equation $\rho = ke^{a(\varphi+\varphi_0)}$, where k and a, constant coefficients,, ϕ_0 it determines the beginning of spiral [3.17].

Of two-pass spiral the second arm is obtained from the first by its shift/shear to 180°. Experimentally detected that with antenna feed in center the intensity of current in arms falls on 20 dB and more after the passage of the turn whose perimeter is approximately equal to wavelength. Therefore a change in the wavelength is equivalent as if rotation of spiral.

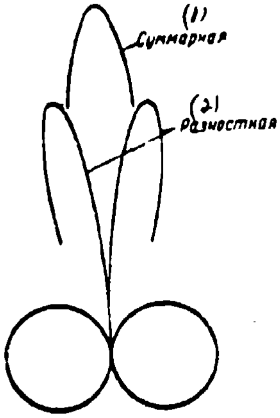

Fig 3.47

Fig. 3.47. Comparison of directional characteristic of circular system with reflector and cosinusoidal characteristic.

Key: (1). Total. (2). Differential.

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The boundary waves of the helical antenna are determined by the length overalls and smallest turns, frequency band reaches 20-fold. To minimum wave besides the diameter of spiral affects the method of antenna feed. Plane spiral can work as electrical and as magnetic (slot) antenna. With the feed of slot antenna to the edges of slot, is soldered the vein/strand and the outer covering of cable. The diameter of cable, thus, limits minimum wavelength. m1 antenna possesses elliptical polarization with an elliptic coefficient of to 2:1. The width of the lug/lobe of radiation pattern at half power varies from 40 to 50°. Traveling-wave ratio with the feeder, which has $\rho_{ф}=50$ ohm, is more than 0.5.

The helical antenna can have a resonator for providing the unidirectional reception. The diameter of resonator must somewhat exceed the diameter of spiral, which is taken equal approximately 0.5 $\lambda_{\text{мин}}$.

The helical antennas are applied mainly at superhigh frequencies. When it is required to obtain in the wide section of ultrashort or short waves the not changing weakly directal radiation pattern and good agreement with feeder, can be used antenna with logarithmic periodic structure (LPA) [3.4, 3.18-3.21].

Diagram of LPA is shown in Fig. 3.49. Antenna consists of two identical parts of I and II. Part II is formed by means of the rotation of part I through 180° around the axis, perpendicular to the plane of antenna and passing through the point of the connection of feeder. The bent on circular arc the teeth of alternating/variable length are the vibrators of antenna. The circular sectors from which are branch/shunted the vibrators, are current distributor. With the location of parts I and II in one plane (Fig. 3.49a) the radiation pattern is obtained bilateral and has major lobes along the axis, perpendicular to the plane of antenna.

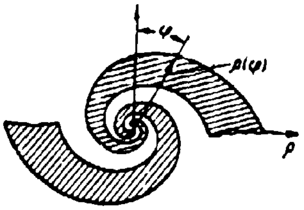

Fig. 3.48. Logarithmic helical antenna.

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If parts I and II are establish/installed at an angle ψ one to another (Fig. 3.49b), then radiation pattern becomes one-sided, moreover its maximum (principal direction) it is directed along the bisector of angle ψ to the side, opposite to the aperture of parts I and II.

The vibrators of LPA can have also trapezoidal form. They are fulfilled from continuous metallic sheet, from the wire, which edges the ducts of antenna, or from single wires. Figures 3.50a depicts LPA with vibrators from single wires at the angle between the parts of antenna $\psi=0$. Single-wire antenna can have also zigzag structure (Fig. 3.50b).

The parameters of LPA are τ, σ, α_1, β_1:

$$\tau=\frac{R_{N+1}}{R_N},\quad \sigma=\frac{r_N}{R_N},$$

where $R_1, R_2, \ldots, R_N$ - distances from in-feed to the periphery edges of vibrators;

$r_1, r_2, \ldots, r_n$ are distances from in-feed to the internal edges of vibrators;

$R_N - r_N$ - thickness of the n vibrator.

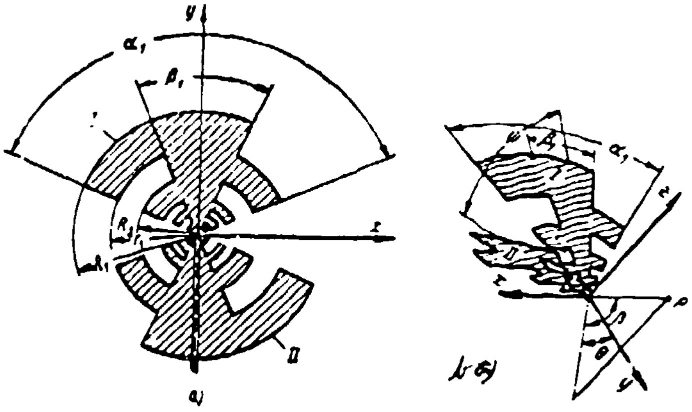

Fig. 3.49. Logarithmic periodic antenna: а) $\psi = 0$; б) $\psi \neq 0$.

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Calculation begins from the vibrator of maximum length.

The constancy of values τ and σ of all vibrators is caused by the structure of antenna.

According to the operating principle of LPA, it is the complex director antenna, which consists of the separate groups of the vibrators each of which includes active vibrator, reflector and director. For the correct work of the group of vibrators, the resistor/resistance of reflector must be inductive, and director - capacitive. This is achieved by the shortening of the length of director and by the elongation of the length of reflector relative to the length active vibrator l_0, inclined to the adopted wave $\lambda (4l_0 = \lambda)$. Then current in reflector anticipate/leads current in active vibrator, and current in director lags behind this current and are satisfied the conditions of maximum reception from director and minimum from reflector.

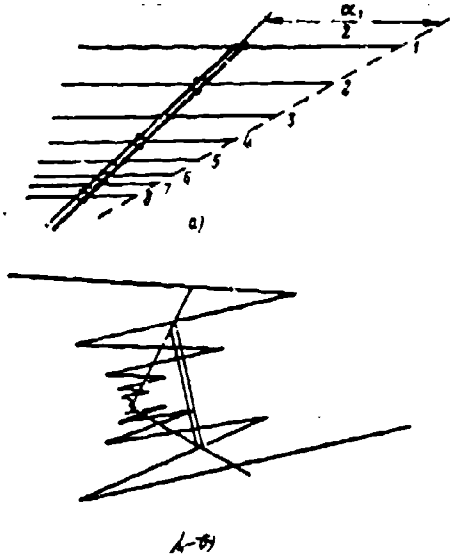

Fig. 3.50. Single-wire logarithmic antenna: a) with single vibrators; b) zigzag.

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Let for the adopted wave (Fig. 3.50) vibrator 4 be inclined into resonance. Vibrator 5 is director, vibrator 3 - by reflector.

Because of the passage of the currents of vibrators 3, 4, 5 along feeder, the voltage drop in common/general/total load from the current of reflector 3 had to lag behind voltage behind the current of active vibrator, and voltage from director's current 5 - to anticipate/lead him. However, because of diagram, obtaining vibrators of the phase of the currents of vibrators 3 and 5 are additionally turned to 180° relative to the current active vibrator. As a result of voltage from the currents of all three vibrators, approximately they coincide in phase and for principal direction are created favorable conditions for the formation of radiation pattern. The currents of the remaining vibrators, strongly detuned in frequency by relatively taken, are low and do not affect the formation of radiation pattern.

During the elongation of wave, comes into action the group of vibrators 2, 3, 4 and of so forth.

With an increase in the angle ψ the three-dimensional/space communication/connection between vibrators decreases, grow/rises the role of current distributor in the establishment of the phases of stresses from vibrators. Simultaneously increases reception by current distributor. To vibrators is realized the reception of horizontal electric field, to current distributor - vertical electric field. Antenna with logarithmic periodic structure possesses therefore circular or elliptical polarization.

If parts I and of II antenna to connect is antiphase, then it is possible to obtain two-lobe radiation pattern with zero in the direction of the bisector of angle ψ between parts.

At each frequency current distributor is loaded by the resistor/resistances of those vibrators which are tuned to frequencies, closest to that which is taken. The reactance of vibrators has different sign and approximately they are compensated for. There remains only effective resistance, and, thus, can be reached good agreement with feeder over a wide range of frequencies.

Frequency band LPA is determined by following considerations.

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On maximum wave $\lambda_{\text{макс}}$ works the longest antenna vibrator by length $l_{\text{макс}}$, moreover $\lambda_{\text{макс}} = 4l_{\text{макс}}$. Respectively for $\lambda_{\text{мин}}$ we have $\lambda_{\text{мин}} = 4l_{\text{мин}}$.

Investigations showed that the vibrator on which is placed more than 2-2.5 waves, significantly makes diagram worse directivity. must be fulfilled condition $l_{макс} < (2 \div 2,5)\lambda_{мин}$. Consequently, relation $\frac{\lambda_{макс}}{\lambda_{мин}}$ the group of vibrators lie/rests within the limits of 8-10 times.

The methods of performance calculation of the directivity of LPA are presented in [3.4, 3.18].

In LPA one should have following relationship/ratios between τ and α_1:

$$\tau = 0,83; \quad 0,8; \quad 0,73; \quad 0,65;$$

$$\alpha_1^0 = 10; \quad 14; \; 19, \; 24, \; 30; \; 37, \; 45.$$

Table 3.4. Some characteristic of LPA.

(1) Параметры			(2) Ширина главного лепестка при уменьшении мощности в 2 раза, *град*		(3) Отношение коэффициентов усиления ЛПА и полуволнового вибратора, *дб*	(4) Максимальный уровень боковых лепестков, *дб*
α_1°	τ	ψ°	(3) при вертикальной поляризации	(4) при горизонтальной поляризации		
75	0,4	30	74	155	3,5	—12,4
75	0,4	60	73	103	5,3	— 8,6
75	0,5	45	67	106	5,6	—14,9
75	0,5	60	68	93	6,1	—12,75
60	0,4	30	85	153	3,0	—12,0
60	0,4	60	87	87	5,3	— 7,0
60	0,5	30	70	118	4,9	—17,7
60	0,5	60	71	77	6,7	— 9,5
60	0,707	45	64	79	7	—15,8
60	0,707	45	66	66	7,7	—12,3

Key: (1). Parameters. (2). Width of main lobe during a decrease in the power 2 times, 2 rad. (3). Ratio of the factors of amplification of LPA and half-wave dipole, dB. (4). Maximum side-lobe level, dB.

Table 3.5. Entry impedances of LPA.

$\psi°$	(1) Волновое сопротивление, ом	(2) Минимальное значение КБВ
60	120	0,7
45	110	0,69
30	105	0,67
7	65	0,55

Key: (1). Wave impedance, ohm. (2). Minimum value of KBV.

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Being assigned τ, α_1 and by the frequency band of the antenna, it is possible to determine its size/dimensions.

Table 3.4 and 3.5 give some characteristics and entry impedance of LPA [3.18].

Use of LPA is possible in radio direction finders with the amplitude and phase-difference methods of the reading of bearing.

3.13. Antennae of the system of radio direction finders at superhigh

frequencies.

In the range of superhigh frequencies from 30 to 300-400 MHz in essence, are accepted the same antenna systems that and at high frequencies (to 30 MHz). Selection by the antenna of system to a considerable degree is determined by the permissible overall sizes of system and by the method of the reading of bearing. In amplitude radio direction finders as by the rotatory antenna of system are utilized the rotatable loop, the diverse framework, the pair of separated antiphase dipoles, connected by H-figurative diagram, or n of the spaced broadside antenna arrays, arrange/located in one plane or of circumference and which create acute/sharp radiation pattern. Rotatable loop is assembled vertically for direction finding on the vertically polarized component of electric field or horizontally for direction finding on horizontal electric field. In the latter case is utilized the doubled framework (see Fig. 3.22). In the construction of the pair of the diverse dipoles at frequencies, large 50 MHz, frequently is provided for the possibility of the simple reorientation of direction of antennas from vertical (Fig. 3.51a) to horizontal (Fig. 3.51b) and vice versa.

The communication/connection of dipole in a pair with input circuit sometimes makes variable within small limits. By the selection of communication/connection at the torque/moment of the reading of bearing they attain obtaining the acute/sharp of zero audibility. For the nondirectional (attendant) reception and determining the single-valued bearing, is applied supplementary central antenna.

Sometimes for determining single-valued bearing, do not provide for supplementary antenna, but is introduced imbalance into circuit of one of the dipoles; the nondirectional reception obtains by means of disconnection/cutoff from diagram of one of the two dipoles.

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For the elimination of the manifestation of resonances in the supporting support into it throw in insulator and both parts of the support connect through resistor/resistance.

If this is required for the selected direction-finding method, is realized the high-spin motion of the pair of antennas by motor, moreover instead of the brushes is applied the special transformer between the antenna and input circuit of receiver.

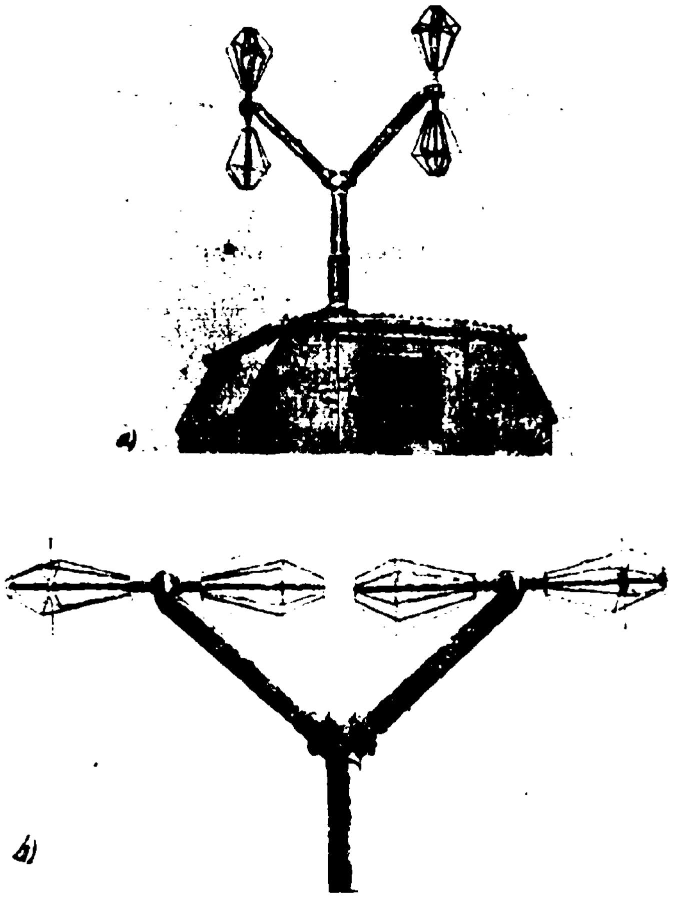

Fig 3.31

Fig. 3.51. UVK of μ-figurative antenna with changing polarization.

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One Winding of transformer, connected to antenna, during rotation by the antenna of system rotates within another, motionless, connected with the input of receiver so that the communication/connection between the antenna and the receiver does not change during rotation. Pair or n cophasal-connected antennas can be utilized with reflector, then immediately is counted off single-valued bearing.

For providing the possibility of direction finding on vertical or on horizontal component of electric field are structurally united those who are connected by switch the pair of antiphase vertical wire antennas and the horizontal dipole whose direction is perpendicular to the plane of the pair of antennas. Horizontal dipole can be applied even one, if direction finding is realized on the horizontal component of electric field.

Is known application/use in the range of frequencies 200-400 MHz of vertical dipole with rotatable around dipole semicylindrical reflector and the reading of bearing according to the principle of the comparison of the phase of the voltage of the frequency of rotation (modulation) from the oriented radio station with the phase of reference voltage. Is utilized also the rotatable V-antenna with reflector which during rotation through 90° is suitable for the direction finding of the vertical and horizontally polarized electric field. Sometimes two separate angle irons, turned n 90°, are establish/installed from two sides of reflector and are changed over by switch of receiver. of simultaneously during switching the indicator of bearing is displaced by 180°.

Are applied goniometric systems with motionless antennas (from 4 to 8) for the directional reception and with central antenna for the nondirectional reception, moreover goniometers are manufactured inductive or capacitive (see §4.7).

For obtaining acute/sharp directional characteristic besides the revolving series a cophasal-connected vibrators, are utilized circular antenna systems with reflector and antenna commutator. The diameter of the arrangement/permutation of antennas reaches (8-10) λ, the number of dipoles in system - to 100 [8.25]. The dipoles of system can be establish/installed at an angle of 45° to vertical line, then is provided direction finding during any polarization of field.

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Direction finding is realized on of the traced on the screen cathode-ray tube to the radiation pattern, which is obtained by rotation by the motor of antenna commutator. In by the circular antenna to system the bearing can be counted off also according to the method of the cyclic measurement of phase in high frequency (see §8.8). The inoperative antennas are closed to the earth. Such systems are applied in ground-based and ship radio direction finders. Since on ultra short waves the fundamental source of errors consists in the effects of the fields of reemitters, increase in the separation is the effective means of an improvement in the accuracy of direction finding.

In order to increase range and to improve the conditions of direction finding (to decrease the errors due to the effect of the near environment), antenna system they raise above the earth's surface and usually they assemble above the metallic counterweight.

For the facilitation of coupling by the antenna of system with their receiver place together on high mast or near antenna they are establish/installed high-frequency ducts and frequency converter, also, on feeders to the receiver, placed below, transmit the voltages of underfrequency. One should consider the possibility of the disturbance in this case of the relationship/ratio of vertical and horizontal fields at close distances from transmitter (see §6.3).

As symmetrical vibrators in by an antenna to system for providing the wide frequency band are utilized cylindrical, conical, basket type or other wide-range dipoles. Are applied also loop-antennas. Frequency band, overlapped by one assembly of dipoles, from 2-3 to 4-5 and more.

If it is required to overlap wider frequency band, sometimes are applied several interchangeable assemblies of dipoles.

In the range of frequencies, greater than 300-400 MHz, in radio direction finders are utilized the horn, parabolic, dielectric, slotted, spiral and other directional antennas of this frequency band. The descriptions of antennas, their characteristics and calculation methods are given in the courses of antennas [8.13, 3.6, 3.5].

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3.14. The multiple operation of the receivers of radio direction finder from oinoi by the antenna of system.

In certain cases it is necessary that two or several receivers of radio direction finder work in parallel from by the common/general/total antenna of system. A parallel connection of receivers can be realized through decoupling resistors which decrease the mutual shunting of receivers and remove the possibility of the penetration of the voltage of the heterodyne of any receiver on the inputs of other receivers.

Figures 3.52 gives the equivalent diagram of parallel connection a of the receivers through the untying ohmic resistances $R_{\text{р}}$.

In figure it is marked:

$E_{\text{а}}$ - the voltage of antenna on the frequency of tuning of one of the receivers;

$R_{\text{а}}$ — antenna resistance;

$R_{\text{вх}1}, R_{\text{вх}2}, \ldots, R_{\text{вх}n}$ - entry impedances of receivers;

Tp - the transformer of the agreement of loads.

We assume that the first receiver is tuned to a frequency of stress E_a and its resistor/resistance $R_{n1}=R_{nx}$, where R_{nx} is entry impedance of receiver during tuning. Remaining receivers are inclined for other frequencies and their entry impedances for a voltage frequency E_a we take equal to zero:

$$R_{n2}=R_{n3}=\ldots=R_{nm}=0.$$

Then most powerful manifests itself by-passing of remaining receivers.

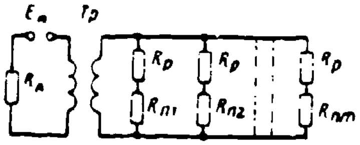

Fig. 3.52. Parallel connection m of receivers into antenna.

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Impedance R_c on secondary winding of transformer under these conditions

$$R_c=\frac{R_p(R_p+R_{вх})}{mR_p+(m-1)R_{вх}}. \qquad (3.99)$$

Knowing R_c and R_a, it is possible to calculate the required transformation ratio of the matching transformer

$$n=\sqrt{\frac{R_a}{R_c}}=\sqrt{\frac{R_a[mR_p+(m-1)R_{вх}]}{R_p(R_p+R_{вх})}}. \qquad (3.100)$$

The coefficient of transmission of direction $E_\text{н}$ to the input of the receiver, tuned to a frequency $E_\text{н}$, in the presence of matching transformer is equal to

$$k_\text{т} = \frac{1}{2}\sqrt{\frac{R_\text{г} R_\text{вх}^2}{R_\text{а}(R_\text{г}+R_\text{вх})[mR_\text{р}+(m-1)R_\text{вх}]}}. \quad (3.101)$$

On the basis of equivalent diagram (Fig. 3.52) the attenuation factor of stress of the heterodyne $U_\text{г}$, which penetrates on the input of receiver, tuned to a frequency of heterodyne, is expressed

$$\gamma_\text{г} = \frac{U_\text{г}}{U'_\text{г}} = \frac{2mR_\text{г}+2(m-1)R_\text{вх}}{R_\text{вх}}. \quad (3.102)$$

In Table 3.6 are designed the values $\frac{k_\text{т}}{\sqrt{\frac{R_\text{вх}}{R_\text{а}}}}$, $\frac{R_\text{о}}{R_\text{вх}}$ and $\gamma_\text{г}$ for different relations $\frac{R_\text{г}}{R_\text{вх}}$ and the different number of receivers. Being assigned $\gamma_\text{г}$, it is possible from table to find $R_\text{р}$, $k_\text{т}$ and n.

Upon the parallel connection of the receivers through decoupling resistors is obtained the loss of stress (3.101). The easiest method of the realization of the decoupling of receivers without the loss of stress is the switching on of the receivers through the separate cathode followers, which are simultaneously the cell/elements of the decoupling between receivers. To the grids of cathode followers, is connected the antenna, into the cathode of each through the feeder, is included the receiver.

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Resistor/resistance in cathode is selected such so that the feeder would be loaded for wave impedance. By a deficiency/lack in the diagram with cathode followers it is characteristic it nonlinear distortions with powerful signals, and also deterioration in the receiver sensitivity due to noise lamp resistance.

For the compensation for the losses of stress between the antenna and the receiver, are applied broadband antenna amplifiers.

During the application/use of an amplifier, just as during the application/use of cathode followers, appear nonlinear distortions at their input. Two voltages with frequencies f_1 and f_2 create the voltage of combination frequency $pf_1 \pm qf_2$. It is necessary so to calculate diagram so that the nonlinear distortions lie/rest within the permissible limits.

For the parallel connection of the receivers of radio direction finder, can be required application/use two- or of three-channel amplifiers. In this case it is necessary to take measures for the achievement of the identity of channels.

Table 3.6. Calculation of the diagram of the decoupling of receivers.

Число приемников m (1)	$R_p=R_{вх}$			$R_p=2R_{вх}$			$R_p=3R_{вх}$			$R_p=4R_{вх}$		
	$\frac{k_т}{\sqrt{\frac{R_{вх}}{R_a}}}$	$\frac{R_c}{R_{вх}}$	γ_r	$\frac{k_т}{\sqrt{\frac{R_{вх}}{R_a}}}$	$\frac{R_c}{R_{вх}}$	γ_r	$\frac{k_т}{\sqrt{\frac{R_{вх}}{R_a}}}$	$\frac{R_c}{R_{вх}}$	γ_r	$\frac{k_т}{\sqrt{\frac{R_{вх}}{R_a}}}$	$\frac{R_c}{R_{вх}}$	γ_r
2	0,204	0,666	6	0,184	1,2	10	0,164	1,72	14	0,145	2,22	18
3	0,158	0,4	10	0,145	0,75	16	0,131	1,09	22	0,12	1,43	28
4	0,134	0,286	14	0,123	0,545	22	0,112	0,8	30	0,108	1,05	38
5	0,118	0,222	18	1,109	0,428	28	0,1	0,642	38	0,091	0,834	48

Key: (1). Number of receivers m.

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Chapter 4.

INSTRUMENT ERRORS.

4.1. Instrument errors of radio direction finder.

As it was shown in chapter 2, instrument errors are caused by deficiency/lacks in the radio direction finder as measuring meter. The tool houses include the errors, characteristic to the operating principle of the selected system, and also production and assembling-adjusting.

Production errors are determined by deficiency/lacks in the structural/design and electrical calculations (by incorrect selection of size/dimensions by the antenna of system, by incorrect calculation of the inductance of the coils of goniometer and so forth), and also by inaccuracies in the production of the separate cell/elements of radio direction finder and radio direction finder as a whole. assembling-adjusting errors are caused by the misadjustment of radio direction finder.

Inaccuracies in production and mounting of radio direction finder sometimes are led to the appearance of the so-called antenna effects which also are related to instrument/tool deficiency/lacks in the radio direction finder.

Are examined below the instrument errors, characteristic to antenna feeder systems. The instrument errors, caused by receiving indicators, are described in chapter 8.

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4.2. Antenna effect within rotatable loop.

The incoming electromagnetic wave creates on the framework certain potential with respect to the earth/ground, because of which appear the bias currents to the earth. If, furthermore, the framework through the connected to it receiver is connected with the earth/ground by conductors, then also in these latter appear emf and currents. The strength of these currents does not depend on the direction of the incoming electromagnetic waves.

Thus, one should distinguish two kinds of the currents, which take place through the turns of the framework: the push-pull current, depending on the direction of the incident wave, and the single-cycle current, which appears between the framework and the earth/ground and not depending on direction of the incident wave. In the latter case the framework operates, as the usual antenna, and this phenomenon obtained the designation of the antenna effect of the framework.

Let the framework be tuned by condenser/capacitor C and to it is connected tube (Fig. 4.1), the end/leads of framework a and b having with respect to the earth/ground capacitance C_1 and C_2. Because of operating in sides CD and ef emf E_1 and E_2 appear the currents:

a) current I_0, which appears around the framework whose value depends on difference E_1 and E_2 and, therefore, on the angle between the plane of propagation and the plane of the framework; this be a fundamental frame current; it creates on condenser/capacitor C the potential difference, which affects the grid of the first tube of amplifier;

b) currents I_1 and I_2, which flow from the framework to the earth through the distributed capacitances of framework C', C'', C''' and of so forth, then through capacitance C_1 and C_2, which are closed again to the framework. Each of these currents is determined only by one of the electromotive forces E_1 and E_2 respectively.

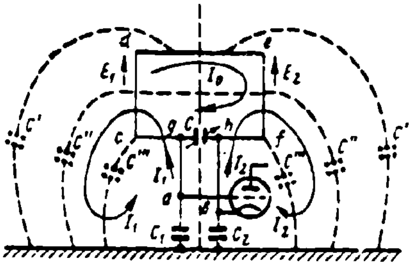

Fig. 4.1. Onset of antenna effects.

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Because of this currents I_1 and I_2 do not depend on the direction of the incident wave. Their radiation pattern is circumference of a similar to the diagram open antenna. These create at points a and b potentials earth referenced:

$$U_a = \frac{I_1}{j\omega C_1} \quad \text{and} \quad \dot{U}_b = \frac{I_2}{j\omega C_2}.$$

The receiver affects a potential difference of points a and b. This potential difference:

$$\dot{U}_b - \dot{U}_a = \frac{I_2}{j\omega C_2} - \frac{I_1}{j\omega C_1}$$

it is turned into zero, if $C_1 = C_2$ and $I_1 = I_2$. Satisfaction of the second condition is possible only in such a case, when the framework is completely symmetrical, i.e., if capacitance relative to the earth/ground of any point c, arrange/located to the left of the axis of symmetry, is equal to the capacitance of symmetrical by it point f. Hence it follows that the antenna effect does not appear in radio direction finder, even if the framework, and the diagram, connected to its end/leads, are completely symmetrical.

The heterogeneity of the capacitance of the framework can be caused by its asymmetric location of the relatively surrounding object/subjects. So, for providing a good work of the contemporary ship radio direction finder, which includes skip band, its frame device is assembled on the top of mast. The currents, induced in mast and within the framework, with the reception of radio station strongly are distinguished. The calculation, given in [10. 3], shows that with the mast with a height/altitude of 16 m and the single-turn framework by the area of 0.5 m² the current in the foundation of mast to 2-3 orders exceeds the current of the directional reception within the framework. During the insignificant displacement of the axis of the framework with respect to the axis of mast due to the manifestation of the dissimilarity of the capacitance of the sides of the framework relative to mast and the surrounding metallic object/subjects into the framework is induced large supplementary nondirectional emf (antenna effect).

Antenna effect in the framework appears also, if, for example, the framework is arrange/located too closely to the wall of building or metallic part of the ship (aircraft) (Fig. 4.2).

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Let us examine the reasons for asymmetry, connected with the circuit diagram of the framework. From the simplest circuit diagram of the framework (Fig. 4.3) it is evident that one end/lead of the framework is connected to the grid of the tube, which has negligible capacitance with respect to the earth/ground, another end/lead is connected to the cathode of tube, which has very considerable capacitance earth referenced. Very frequently the cathode of tube directly is grounded, and then asymmetry will be expressed still more powerful.

If symmetry conditions are not observed and is observed antenna effect, then the resulting radiation pattern is distorted, he differs from the studied in us ideal diagram. To the grid of tube, will be supplied two voltages: depending on the direction of incilent wave and not depending on this direction.

Taking into account the aforesaid, we can use conclusions §3.8, since the voltages, which are received because of the antenna effect of the framework, are completely analogous to the voltage of the open antenna with the combined reception on the antenna and the framework. Under normal conditions the grid voltage of tube, which is received due to antenna effect, is lower than the voltage from the framework. It is analogous with that, as we enter §3.8, examining the action of the open antenna, it is possible the voltage of antenna affect to decompose on two those who compose: one in phase, another - out of phase in $\pi/2$ relative to the voltage, which is the result of the action of the pure/clean framework.

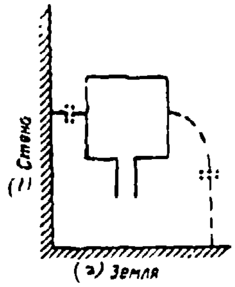

FIG. 4.2

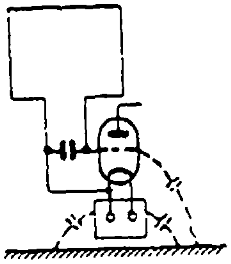

Fig 4.3

Fig. 4.2. Asymmetry of framework of relatively surrounding object/subjects.

Key: (1). Wall. (2). Earth.

Fig. 4.3. Asymmetric diagram of framework.

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There shown, that:

- phase component/term causes the displacement of the minimums; the minimums of reception, distant one from another into the ideal diagram of reception accurately to 180°, into this case they are mutually located on smaller angle;

- nonphase component it causes diffuseness, the vagueness of the minimums of intensity of reception. If with the ideal diagram of reception in the position of the framework, perpendicular to the direction of propagation of wave, completely there is no reception, then under nonphase antenna effect is obtained only the minimum of reception, by those more diffuse, than is more antenna effect.

Figures 4.4 depicts the diagram of reception with the existence of antenna effect in phase with frame. The directions, by which the intensity of reception is equal to zero, differ by angle ±Δ from the directions of zero reception for a purely frame diagram. Counting off bearing with the aid of the direction finder, which possesses this diagram of reception, we complete the error, equal to +Δ or Δ depending on that, on which of two minimums is counted off the bearing. This error cannot be taken into account by calibration, since it depends on the wavelength and direction of incoming electromagnetic field. Its effect can be excluded, if we orient first on one minimum, and then, after turning the framework to angle of 180 ± 2Δ, on another. Calling the first bearing q_1, and the second q_2, we will obtain the true bearing from the formula

$$q = \frac{q_1 + (q_2 - 180^\circ)}{2}.$$

Into normally working direction finder the displacement of the minimums, or as it is accepted to call this phenomenon, the fracture of the axis of the minimums, must no or be very small (for example, the fracture of the axis of the minimums $2\Delta < 1^\circ - 1.5^\circ$).

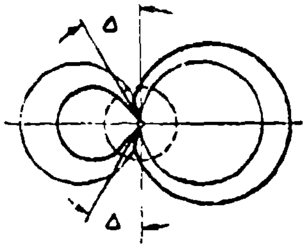

Fig. 4.4. Radiation pattern under antenna effect in phase with frame.

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Figures 4.5 depicts radiation pattern in the presence of nonphase antenna effect, and also vector diagram of addition emf. Here the minimums retain their direction, but instead of the complete zero receptions, are obtained the diffuse, little determined minimums. This diffuseness is led to the appearance of subjective errors of reading.

Actually, the total voltage will be in this case equally

$$U=\sqrt{U_{р}^{2}\sin^{2}\theta+U_{з}^{2}},$$

where $U_{\text{р}}$ - is voltage across capacitor from frame emf;

$U_{\text{а}}$ - voltage from antenna effect;

θ is an angle of rotation of the framework from the direction of zero reception.

In the position of the minimum, the voltage is turned in $U_{\text{мин}} = U_{\text{а}}$.

Let us assume that this stress higher than equivalent disturbing voltage and noises determines the angle of the equal to audibility. An increment in the audibility during the rotation of the framework from θ = 0 through small angle Δθ will be

$$\Delta L = 20\lg\sqrt{\frac{U_{\text{р}}^2\sin^2\Delta\theta + U_{\text{а}}^2}{U_{\text{а}}^2}} = 10\lg\left[1 + \frac{U_{\text{р}}^2}{U_{\text{а}}^2}\sin^2\Delta\theta\right].$$

With small Δθ, applying the formula of approximation calculus, we obtain

$$\Delta L = 4{,}34\,\frac{U_{\text{р}}^2}{U_{\text{а}}^2}(\Delta\theta)^2$$

and

$$\Delta\theta = 0{,}48\sqrt{\Delta L}\,\frac{U_{\text{а}}}{U_{\text{р}}}.$$

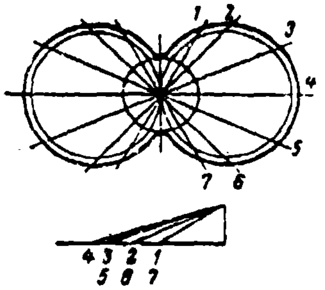

Fig. 4.5. Radiation pattern under nonphase antenna effect; vector diagram of addition emf.

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For example, limiting angle by equal to audibility $2\Delta\theta$ value 3° ≈ 0.05 rad and considering $\Delta l = 1$ dB, we obtain the following condition, which limits the value of the antenna effect:

$$\frac{U_{\text{а}}}{U_{\text{г}}} < \frac{0{,}05}{2\cdot 0{,}48\sqrt{1}} \approx 0{,}05.$$

Hence it is apparent that even the relatively low value of antenna effect produces an essential increase in the subjective error. Thus, for obtaining precision work of direction fiader must no the extra-phase component of antenna effect. ¶ it must be noted that usually the phase of emf of the antenna effect when the size/dimensions of the framework are much less than the wavelength, differs from the frame by the angle, close to $\pi/2$, since the system of the framework, considered as open antenna, is not adjusted.

4.3. Elimination of antenna effect within rotatable loop.

Since the reasons for antenna effect is the asymmetry of the framework or connected to it receiver, and also reception by an entire system of the framework as by open antenna, the measures of the elimination of antenna effect are reduced to obtaining of the completely balanced network of the switching on of the framework and to the exception/elimination of reception by the framework as open antenna.

Let us examine the diagrams, which ensure the symmetry of the switching on of the framework. Protozoan of these diagrams - the grounding of midpoint is depicted on Fig. 4.6. Grounding of midpoint will entail the need for the connection of receiver to the half of the framework: one end/lead of framework a is connected to grid, another b remains free. The cathode of tube is connected with midpoint of framework O. Because of very small grid capacitance earth referenced the difference in the capacitance of end/leads a and b is so small that the antenna effect appears to very weak degree.

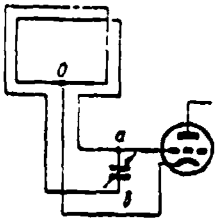

Fig. 4.6. Balanced network of connection of receiver to framework.

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Furthermore, induced in ground wire emf are directed opposite to emf, induced in lead wires, which also leads to a decrease in the antenna effect.

Second method - switching on at the input of two tubes according to push-pull circuit (Fig. 4.7). Here both end/leads of the framework are connected to the grids of two tubes, which must ensure the symmetry of diagram.

Finally, symmetry can be restore/reduced by the connection of the special balancing condenser/capacitor to the end/leads of the framework, which have different capacitance earth referenced. But if, as so often in the case, the cathode of the tube directly is grounded, then this method cannot be applied, since the second end/lead of the framework we ground cannot. Therefore the method in question they always combine with diagram, it is depicted on Fig. 4.6.

In practical fulfillment the diagram takes the form, shown in Fig. 4.8. For larger convenience is here used the differential capacitor, which makes it possible to increase capacitance earth referenced any end/lead of the framework. A deficiency/lack in this method is the effect of the compensating capacitance on the tuning of the framework.

Decrease in the asymmetry of diagram and value of antenna effect gives to known degree of application of the diagram of the inductive coupling of the framework (Fig. 4.9). Results are obtained still better, if midpoint of coupling coil L_1 is grounded.

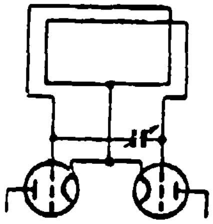

Fig.4.7.

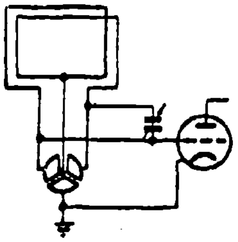

Fig.4.8.

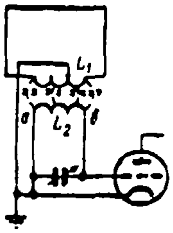

Fig.4.9.

Fig. 4.7. Push-pull circuit of switching on of framework.

Fig. 4.8. Diagram of compensation for antenna effect.

Fig. 4.9. Inductive circuit diagram of framework.

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It is possible to obtain a further decrease in the antenna effect, if we include/connect the tuned circuit symmetrically, for example, after connecting tube to the half of coil L_2 (as in Fig. 4.6).

With the powerful communication/connection between coils L_1 and L_2 because of an increase also the capacitance of one coil relative to another appear the bias currents from L_1 and L_2. If the switching on of coil L_2 not is completely symmetrical, then the distribution of these ticks along coil will be nonuniform. Thus, for instance, in Fig. 4.9 bias current of the end/lead of coil a will be more than in the end/lead of coil b, since the first path represents smaller resistor/resistance to the earth/ground. the inequality of these currents produces equalizing currents of antenna effect in L_1.

The effect of bias currents can be considerably attenuate/weakened by the application/use of the electrostatic shield between coils L_1 and L_2. The electrostatic shield is the metallic separator, connected with the earth/ground. It is better, if it is comprised from separate conductors. Electrical lines of force cannot intersect the conductor of screen and therefore they do not pass from one coil to another. Magnetic induction remains almost without change, since screen must be extended and be arrange/located so that in it uld not arise the circular demagnitizing currents.

Were proposed the special diagrams of the inductive connection, reducing the effect of the asymmetry of input circuit of radio direction finder on the framework.

Figures 4.10 shows diagram with the transformer, proposed by V. D. Kuznetsov. Secondary winding of transformer, connected in the symmetrical framework, consists of two halves L_2' and L_2'', wound tightly one to another and to primary winding L_1. The common point of halves O is grounded.

Because of this winding/coil the mutual inductance of the half of secondary winding ωM_2 is approximately equal to their inductance, i.e.,

$$\omega M_2 \approx \omega L'_2 = \omega L''_2.$$

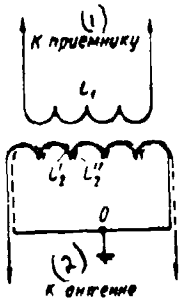

Fig. 4.10. Circuit of V. D. Kuznetsov's transformer.

Key: (1). To receiver. (2). To antenna.

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Let the capacitance earth referenced any point of primary winding (with unsymmetrical loading) cause in one of the halves of secondary winding current I_2 the voltage drop earth referenced $E'_2 = j\omega L_2' I_2$. Because of the mutual inductance of the halves of secondary winding in the second half, will appear the voltage earth referenced $E''_2 = j\omega M_2 I_2$, in value equal to E'_2 and on phase differing by 180°. Both voltages E'_2 and E''_2 average out and will not cause asymmetry in diagram. Therefore antenna effect does not appear. By the selection of cell/elements and by the selection of diagram it is possible to attain the work of transformer in broadband [3.4].

Figures 4.11 shows diagram with the transformer, proposed by A. A. Pistol'kors. Winding of the transformer, connected in the directional antenna, is wound in the form of two halves L'_3 and L''_3 between which is located the winding of input circuit. The latter consists of two identical coils L_1 and L_2, wound towards and connected in parallel. The common points of windings L_1 and L_2 are connected: point O - with the grid of the first tube of receiver, point O' - with the earth/ground.

As a result of this winding/coil of transformer calling asymmetry single-cycle coil current, connected to receiver, does not manifest itself the coil current, connected to the directional antenna.

We will examine the circuits of the elimination of single-cycle current in loop antenna. Asymmetry of receiver circuit produces the appearance of a single-cycle current and antenna effect in any symmetrical directed antenna. is removed single-cycle current by the selection of the circuit diagram of the symmetrical antenna, analogous examined for a rotatable loop.

The exception/elimination of reception by the framework as open antenna due to the asymmetry of diagram of input can be obtained by the appropriate screening of the framework.

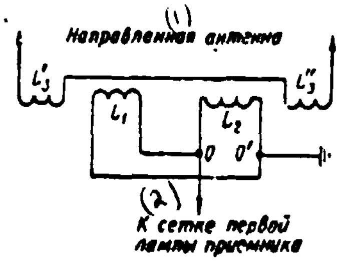

Fig. 4.11. Circuit of A. A. Pistol'kors's transformer.

Key: (1). Directional antenna. (2). To the grid of the first tube of receiver.

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The diagram of this screening is shown in Fig. 4.12: the screen between points A and B is torn and in it does not appear ring currents on the duct of the framework.

Since the reception is realized because of the voltages, induced with field in screen, asymmetry of diagram cannot be the reason for antenna effects.

It is necessary, however, to keep in mind that the asymmetry of the screen of the relatively surrounding object/subjects produces the dissymmetry of currents in it, which leads to the appearance of an antenna effect within the framework. Of the symmetrical doubled framework, examined into § 3.7, emf, which appear in it due to capacitive coupling earth referenced of the right or left side of the framework, are compensated for and therefore the dissimilarity of the capacitance of the sides of the relatively surrounding metallic object/subjects does not manifest itself (see Fig. 3.22).

The methods of symmetrization during the use of coaxial cable for the connection of symmetrical antenna are described in § 4.12.

4.4. Instrument errors of system with rotatable loop.

Most idle time in the relation to diagram and construction is the system of the revolving framework. It is natural that in this system the number of sources of the instrument errors is smallest.

Lateral effect of the framework. In the case of the three-dimensional/space framework, the wire of the framework during the series circuit of its turns forms one supplementary turn, perpendicular to the fundamental turns of the framework. The presence of this turn produces the error in the determination of bearing whose value was determined (to 3.24):

$$\Delta = \operatorname{arc\,tg} \frac{S_{\text{н}}}{S_N}.$$

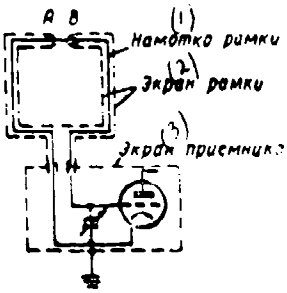

Fig. 4.12. Screening of the framework.

Key: (1). Winding/coil framework. (2). Screen of the framework. (3). Screen of receiver.

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This error has the constant value, which does not depend on wavelength. Therefore it easily can be taken into account during the calibration of direction finder. For a decrease in the lateral effect of the framework, it is possible to fulfill the winding of the framework so that its each turn lie/rests strictly at one plane, and the connections between turns represent sections of wires, perpendicular to the plane of turn.

Effect of bias currents. In the case of the three-dimensional/space framework, is observed also the appearance of the diffuse minimums, even if there is no antenna effect. is explained this phenomenon to the facts that when the three-dimensional/space framework is located in the position, which corresponds to the minimum of reception, its separate turns are found under different potentials. Because of this between separate turns, appear the bias currents through the capacitance of one turn with respect to another. Value of these currents is changed according to cosinusoidal law during the rotation of the framework. Since for these currents circuit is detuned, they are out of phase in w/2 relative to the fundamental current of the framework and, therefore, is produced the diffuseness of the minimum, without displacing it. In this respect they are similar to the nonphase component of antenna effect. Bias currents differ from nonphase antenna effect in the facts that the phase of bias currents is changed by reverse/inverse during the rotation of the framework through 180°, while value and the phase of antenna effect do not depend on the rotation of the framework.

Figure 4.13 depicts radiation pattern and vector diagram of its construction upon consideration of bias currents. Two Large contacted circumferences A represent the diagram of the reception of the ideal framework.

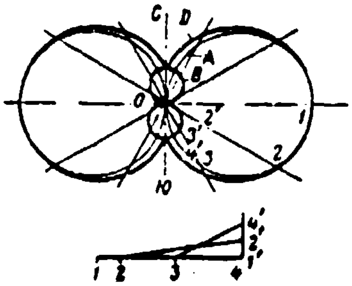

Fig. 4.13. Radiation pattern when bias currents are present,. Vector diagram of addition emf.

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The effect of bias currents is characterized by two contacted circumferences B less in value and shifted to 90° relative to the diagram ideal framework. Diagram D gives resulting heart-shaped diagram. The effect of bias currents just as lateral effect, is absent at the flat/plane framework.

Direct reception. Reception for the input/introductions of the framework, coil and wire receiver creates supplementary emf whose interaction with the fundamental of emf of the framework produces either error or the diffuse minimums depending on the phase of these supplementary emf. If supplementary emf are induced in the stationary parts of direction finder, then their effect completely is analogous with the antenna effect: phase components will produce the displacement of the minimums (i.e. the errorwhose sign is changed during the rotation of the framework through 180°), and nonphase - diffuseness of the minimums, the degree of diffuseness not depending on the angle of rotation of the framework. But if supplementary emf are induced in the parts, which rotate together with the framework (for example, in its input/introductions), then phase components will produce the error whose sign is not changed during the rotation of framework through 180° a nonphase - the diffuseness of the minimum, analogous to the effect of bias currents.

For the elimination of direct reception, it is necessary to ensure the careful screening of an entire direction-finding installation (input/introductions of the framework and receiver). The effective screening of input/introductions considerably is facilitated, if the framework is also shielded.

In supply leads, also they can be induced by emf of direct reception. So that these emf do not affect the circuit of receiver, it is necessary to use decoupling filters in feed circuits.

Reradiation of receiver. Circuital currents of receiver can give reradiation. Secondary field affects the framework and is the source of the errors or diffuse minimums depending on the relationship/ratio of its phase with the phase of ground field.

Special danger represent the last/latter cascade/stages of amplification in fundamental frequency, since in their ducts circulate the considerably amplified currents, capable of creating the sufficiently intense fields of reradiation. For the exception/elimination of this source of errors here, as in the case of direct reception, is necessary the very careful screening of receiver.

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Slope/inclination of the axis of the framework. The slope/inclination of the axis of the framework creates the error whose maximum value can be determined by the formula

$$\Delta = \frac{\gamma^2}{229}, \overset{(1)}{\text{град}},$$

Key: (1). deg.

where γ is angle of the slope of the axis of the framework, deg.

In ground-based installations this error can be made sufficiently small. It is a different matter aboard the ship or aircraft with bank; for example, with bank into 30° error will be approximately 4°.

Eccentricity of the scale. Eccentricity of the scale relative to the rotational axis of the framework produces the error

$$\Delta=\left(\frac{e_1}{\rho}\cos\theta+\frac{e_2}{\rho}\sin\theta\right)57{,}3\ \overset{(1)}{\text{град}}.$$

Key: (1). deg.

where e_1 is eccentricity in direction 0-180°; e_2 eccentricity in direction 90-270°; ρ is a radius of the scale.

Incorrect engraving of the scale, nonuniform divisions, etc,.

Inaccurate zero-setting scales.

Effect of the nonparallelism of the framework in the system of the diverse framework. Besides errors which are inherent rotatable loop, in the system diverse framework can manifest itself their nonparallelism.

The cases of the nonparallelism of the framework in the system of the diverse framework are depicted on Fig. 4.14.

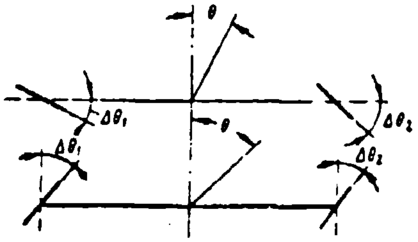

Fig. 4.14. Not in parallel diverse framework.

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Here $\Delta\theta_1$ and of $\Delta\theta_2$ - the angles of deflection of the planes of the framework from their regular arrangement.

In the expressions for emf, induced within the separate framework, instead of the angles θ will enter the angles $\theta + \Delta\theta_1$ and $\theta + \Delta\theta_2$. Since $\cos(\theta + \Delta\theta) \approx \cos\theta - \Delta\theta \sin\theta$ and $\sin(\theta + \Delta\theta) \approx \sin\theta + \Delta\theta \cos\theta$, to each of the diverse framework seemingly had had added the supplementary, perpendicular to it framework with the relative effective height of $\Delta\theta_1$ radian and $\Delta\theta_2$ radian. To the radiation pattern of basic system, will be applied the radiation pattern of the diverse framework, perpendicular to fundamental, emf of the supplementary framework can store/add up not differentially, but accordingly. Then the action of the supplementary framework stops of equivalently to the action simple rotatable loop. furthermore, if $\Delta\theta_1 \neq \Delta\theta_2$, then will appear the antenna effect, proportional to the angle of the nonparallelism of the framework $(\Delta\theta_1 - \Delta\theta_2)$.

For the limitation of the indicated ill effects (maximum error 30') the diverse framework must be parallel with accuracy 10'-15'.

4.5. Antenna effect in a goniometric system.

As within the simple revolving framework, in goniometric system can be observed phase, so also nonphase antenna effect.

Let us examine simplest system of two mutually perpendicular framework (or two pairs of the spaced antennas).

Let us designate: a_1, a_2 are ratio of the amplitudes of emf of the phase antenna effects of the framework of goniometric system to the amplitude of maximum emf of the directional reception; b_1, b_2 - the ratio of the amplitudes of emf of the nonphase antenna effects of the framework to the amplitude of maximum emf of the directional reception $E \cos \theta$ and $E \sin \theta$ — emf, induced within the framework.

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Total emf within the framework taking into account antenna effects are expressed:

$$\dot{E}_1 = E\,[(\cos\theta + a_1) + jb_1],$$
$$\dot{E}_2 = E\,[(\sin\theta + a_2) + jb_2].$$

In the general case the resulting magnetic field in goniometer will be elliptical. For determining the angle of bearing $\alpha_{\text{мин}}$ (direction of the transverse of field) and the relation of small and semimajor axes of magnetic field (A/B) we will use formulas (III.3) and (III.7) of appendix III, after substituting in them values of l = $\cos\theta + a_1$, $m = b_1$, $-n = \sin\theta + a_2$, $-p = b_2$. Then we obtain:

$$\operatorname{tg} 2\alpha_{\text{мин}} = \frac{2\,[(a_1a_2 + b_1b_2) + a_1\sin\theta + a_2\cos\theta] + \sin 2\theta}{(a_1^2 + b_1^2) - (a_2^2 + b_2^2) + 2\,(a_1\cos\theta - a_2\sin\theta) + \cos 2\theta}, \quad (4.1)$$

$$\frac{\frac{A}{B}}{1 - \left(\frac{A}{B}\right)^2} = \frac{[(a_2b_1 - a_1b_2) + b_1\sin\theta - b_2\cos\theta]}{[(a_1a_2 + b_1b_2) + a_1\sin\theta + a_2\cos\theta + \sin 2\theta]}\sin 2\alpha_{\text{мин}}. \quad (4.2)$$

From (4.1) it follows that the bearing error Δ in the general case is designed from the formula

$$\operatorname{tg} 2\Delta \approx 2\Delta = \operatorname{tg}(2\alpha_{\text{мин}} - 2\theta) =$$
$$= \frac{D\cos 2\theta - E\sin 2\theta + 2a_2\cos\theta - 2a_1\sin\theta}{E\cos 2\theta + D\sin 2\theta + 2a_1\cos\theta + 2a_2\sin\theta + 1}, \quad (4.3)$$

where

and

$$D = 2(a_1a_2 + b_1b_2) \text{ и } E = (a_1^2 + b_1^2) - (a_2^2 + b_2^2)$$

are error coefficients.

Let us examine special cases of the presence of antenna effect in one of the framework.

1. $b_1 = a_2 = b_2 = 0$; $a_1 \neq 0$, i.e., only at the first framework has phase antenna effect.

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From (4.3) it follows that then

$$\Delta \approx -0{,}5 \frac{2a_1 \sin\theta + a_1^2 \sin 2\theta}{1 + 2a_1 \cos\theta + a_1^2 \cos 2\theta}$$

with approximately with $a_1 \ll 1$

Formula (4.4) shows that in this case the error changes sign

$$\Delta = -\frac{a_1 \sin\theta}{1 + 2a_1 \cos\theta}. \qquad (4.4)$$

during a change in the angle θ to 180°, i.e., occurs the fracture of the axis of the minimums, equal to 2Δ. after equating zero derivative of (4.4) in terms of θ, let us find the value $\theta_{\text{макс}}$. at which is obtained the maximum error:

$$\cos\theta_{\text{макс}} = -2a_1,$$
$$\theta_{\text{макс}} = \arccos(-2a_1).$$

Maximum error will be

$$\Delta_{\text{макс}} = \frac{a_1}{\sqrt{1-4a_1^2}}. \qquad (4.4')$$

Ratio A/B in this case is equal to 0, i.e., magnetic field in goniometer linear sinusoidal.

If we restrict the fracture of the axis of the minimums by value 2° (Δ = 1°), then phase antenna effect must be not more than by 1.8o/o maximum emf of the framework (or the pair of the spaced antennas).

2. $a_1 = a_2 = b_2 = 0$; $b_1 \neq 0$, i.e., at first framework has nonphase antenna effect.

From (4.3) it follows

$$\Delta \approx -0{,}5\frac{b_1^2 \sin 2\theta}{1+b_1^2\cos 2\theta}. \qquad (4.5)$$

Let us find the value $\theta_{\text{макс}}$. by which is obtained maximum error, after equating derivative of (4.5) in terms of θ zero:

$$\cos 2\theta_{\text{макс}} = -b_1^2.$$

After substituting value $\theta_{\text{макс}}$ in (4.5), we will obtain for the maximum error

$$\Delta_{\text{макс}} = -0{,}5\frac{b_1^2}{\sqrt{1-b_1^4}}. \qquad (4.5')$$

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Expression for the relation of the semi-axes of the ellipse of field on the basis (4.2) takes the form

$$\frac{2\frac{A}{B}}{1-\left(\frac{A}{B}\right)^2} = \frac{2b_1\sin\theta}{\sqrt{1+2b_1^2\cos 2\theta + b_1^4}} \approx \frac{2b_1\sin\theta}{1+b_1^2\cos 2\theta}.$$

For the low values b_1, we obtain approximately

$$\frac{2\frac{A}{B}}{1-\left(\frac{A}{B}\right)^2} \approx 2b_1\sin\theta \;\; \overset{\text{and}}{\text{и}} \;\; \left(\frac{A}{B}\right)_{\text{макс}} \approx b_1.$$

If we restrict error by value 30°, then the permissible nonphase antenna effect must comprise not more than 10-12o/o of frame reception. in this case, the ellipticity of magnetic field (relation of small and semimajor axes) will be also A/B ≈ 10-12o/o.

From the examined special cases escape/ensue those high requirements which are presented to goniometric systems in the relation to the limitation of antenna effects. These requirements are led to the need for the fulfillment of the symmetry of the diagram of framework and pairs of the spaced antennas, and also the complete identity of the parameters of the spaced antennas.

The asymmetry, caused by input circuit, in goniometric systems is attenuate/weakened by the application/use of inductive coupling (in goniometer).

For further decrease in the antenna effect is applied the electrostatic shield in goniometer or supplementary intervening transformer with the screen between windings. The grounding of midpoint of system is also the effective measure of a decrease in the antenna effect.

The limitations of antenna effect, determined for a system of two pairs of the spaced antennas, approximately are retained for a system from n of the spaced antennas.

Reasons for appearance of antenna effects of systems with the spaced antennas are examined in § 4.8.

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Errors of goniometric system.

Fundamental reasons for the disturbance of the goniometric system: the dissimilarity of maximum currents in the field coils of goniometer, communication/connection between the field coils of goniometer or between the framework.

1. Dissimilarity of maximum currents in field coils of goniometer is caused by facts that are different resistor/resistances of ducts of framework (pairs of spaced antennas), maximum mutual inductance of search coil with field coils of goniometer, either maximum emf within the framework, i.e., $Z_{11} = Z_{22}$ or $M_{1\,\text{макс}} \neq M_{2\,\text{макс}}$, or $E_{1\,\text{макс}} \neq E_{2\,\text{макс}}$. the remaining conditions, which ensure the normal operation of system (3.46), are satisfied.

Let us examine the effect of inequality on module/modulus and argument of the resistor/resistance of the ducts of the framework.

Let us accept:

$$\frac{Z_{11}}{Z_{22}} = ae^{j\varphi} = a\cos\varphi + ja\sin\varphi, \qquad (4.6)$$

where $z_{11}/z_{22} = a$ - the relation of moduli of resistance; φ is a difference in their arguments.

After substituting (4.6) and (3.46) with the exception $Z_{11} = Z_{22}$ in (3.44), we will obtain

$$I_3 = j\omega M E h_{a\phi} \frac{\sin\alpha\cos\theta - \frac{Z_{11}}{Z_{22}}\cos\alpha\sin\theta}{Z_{11}Z_{22} - \frac{Z_{11}}{Z_{22}}\omega^2 M^2\cos^2\alpha - \omega^2 M^2\sin^2\alpha},$$

$$I_3 = j\frac{\omega M}{Z_{11}Z_{22}} E h_{a\psi} \times$$

$$\times \frac{\sin\alpha\cos\theta - a\cos\varphi\cos\alpha\sin\theta - ja\sin\varphi\cos\alpha\sin\theta}{1 - a\frac{\omega^2 M^2}{Z_{11}Z_{22}}\cos\varphi\cos^2\alpha - \frac{\omega^2 M^2}{Z_{11}Z_{22}}\sin^2\alpha - ja\frac{\omega^2 M^2}{Z_{11}Z_{22}}\sin\varphi\cos^2\alpha}.$$

Approximately it is possible to count which at the small φ and values a, close to unity, denominator barely depends on α and for current I_3 it is possible to write the expression

$$I_3 \approx j\frac{\omega M}{Z_{11}Z_{22}} E h_{a\phi} \frac{-(a\cos\varphi\sin\theta + ja\sin\varphi\sin\theta)\cos\alpha + \cos\theta\sin\alpha}{1 - \frac{\omega^2 M^2}{Z_{11}Z_{22}}}. \quad (4.7)$$

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After reject/throwing the terms, independent of α, θ and φ, we will obtain

$$I_3 = \cos\theta\sin\alpha - (a\sin\theta\cos\varphi + ja\sin\theta\sin\varphi)\cos\alpha = L \quad (4.8)$$

or

$$I_3^2 \doteq a^2\sin^2\theta\cos^2\alpha+\cos^2\theta\sin^2\alpha-2a\sin\theta\cos\theta\sin\alpha\cos\alpha\cos\varphi, \tag{4.8'}$$

where $\doteq$ - proportionality sign.

Expression (4.8) corresponds to the equation of ellipse. Since current in search coil I_3 is proportional to its penetrating magnetic flux, of (4.8) it follows that the magnetic field in goniometer has the elliptical law of change depending on angle α. During the rotation of search coil never will completely disappear the audibility. When the plane of search coil is perpendicular to the minor axis of the ellipse of magnetic field, is obtained the minimum of audibility; when the plane of search coil is perpendicular to the transverse, is obtained the maximum of audibility.

Let us use formulas (III.3) and (III.7) of appendix III to the solution to equation (4.8). For this, we assume that

$$l=-a\sin\theta\cos\varphi,\ m=-a\sin\theta\sin\varphi,\ -n=\cos\theta,\ p=0.$$

then

$$\operatorname{tg}2\alpha_{\text{мин}}=\frac{2a\sin\theta\cos\theta\cos\varphi}{-a^2\sin^2\theta(\cos^2\varphi+\sin^2\varphi)+\cos^2\theta}=$$

$$=\frac{a\sin2\theta\cos\varphi}{\cos^2\theta-a^2\sin^2\theta}=\frac{2a\operatorname{tg}\theta\cos\varphi}{1-a^2\operatorname{tg}^2\theta}=\operatorname{tg}2\theta_1\cos\varphi, \tag{4.9}$$

where

$$\operatorname{tg}\theta_1 = a\operatorname{tg}\theta;$$

$$\frac{2\frac{A}{B}}{1+\left(\frac{A}{B}\right)^2} = \frac{-a\sin\theta\cos\theta\sin\varphi}{-a\sin\theta\cos\theta\cos\varphi}\sin 2\alpha_{\text{мин}} = \operatorname{tg}\varphi\sin 2\alpha_{\text{мин}}. \tag{4.10}$$

If we determine bearing by the minimum of audibility, then will be obtained the bearing error $\Delta = \alpha_{\text{мин}} - \theta$, which can be found from the expression

$$\operatorname{tg}2\Delta = \operatorname{tg}2(\alpha_{\text{мин}} - \theta) =$$
$$= \frac{2(a^2-1)\sin 2\theta - (1-2a\cos\varphi+a^2)\sin 4\theta}{(1+2a\cos\varphi+a^2) - 2(a^2-1)\cos 2\theta + (1-2a\cos\varphi+a^2)\cos 4\theta}. \tag{4.11}$$

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Let us designate

$$D = \frac{a^2-1}{1+2a\cos\varphi+a^2} \quad \overset{\text{and}}{\text{и}} \quad K = \frac{1-2a\cos\varphi+a^2}{1+2a\cos\varphi+a^2}. \tag{4.12}$$

From (4.11) we will obtain

$$\operatorname{tg}2\Delta = \frac{2D\sin 2\theta - K\sin 4\theta}{1-2D\cos 2\theta + K\cos 4\theta}. \tag{4.13}$$

the maximum value of error let us find further.

From expression (4.10), it follows that $\frac{2\frac{A}{B}}{1-\left(\frac{A}{B}\right)^2}$ it reaches the maximum value when $a_{\text{мин}}=45°$. This, as can be seen from (4.9), it corresponds to the condition

$$\operatorname{ctg} 0_{\left(\frac{A}{B}\right)_{\text{макс}}}=a$$

independent of the value of a phase difference φ.

Let us substitute $a_{\text{мин}}=45°$ in (4.10) and we will obtain

$$\frac{2\left(\frac{A}{B}\right)_{\text{макс}}}{1-\left(\frac{A}{B}\right)^2_{\text{макс}}}=\operatorname{tg}\varphi \text{ или } \left(\frac{A}{B}\right)_{\text{макс}}=\operatorname{tg}\frac{\varphi}{2}$$

independent of value of a.

At the low values of a phase difference φ and with value a, close to unity, from (4.10) it follows that

$$\frac{A}{B}=\frac{\varphi}{2}\sin 20 \text{ deg.}$$

and at θ = 45° we obtain $\left(\frac{A}{B}\right)_{\text{макс}}=\frac{\varphi}{2} \text{ deg}$

We will examine the effect of the dissimilarity of the resistor/resistances of the circuits of the framework. Analogous action will produce inequality on amplitude and phase of the maximum of emf within the framework. In this case of a and ϕ the preceding/previous conclusions must be related to $E_{1\,\text{макс}}$ and $E_{2\,\text{макс}}$.

The obtained expressions (4.9)-(4.13) characterize the error and the diffuseness of bearings in the general case of the presence of dissimilarity in the ducts of the framework.

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When $\phi = 0$ and $a \neq 1$, i.e., the arguments of the resistor/resistances of the framework are identical, and moduli of resistance are different, from (4.9), (4.11) and (4.12) it follows that:

$$\operatorname{tg} \alpha_{\text{мин}} = a \operatorname{tg} \theta, \qquad (4.14)$$

$$D_1 = \frac{a^2-1}{1+2a+a^2} = \frac{a-1}{a+1}, \quad K_1 = \frac{1-2a+a^2}{1+2a+a^2} = D_1^2$$

and

$$\operatorname{tg} 2\Delta_1 = \frac{2\frac{a-1}{a+1}\sin 2\theta - \left(\frac{a-1}{a+1}\right)^2 \sin 4\theta}{1 - 2\frac{a-1}{a+1}\cos 2\theta + \left(\frac{a-1}{a+1}\right)^2 \cos 4\theta}. \qquad (4.15)$$

For this case the error can be found directly from expression for current I_3 (3.44):

$$I_3 = -\frac{j\omega M E\, h_{\text{дф}}\,(Z_{22}\sin\alpha\cos\theta - Z_{11}\cos\alpha\sin\theta)}{Z_{11} Z_{22}\left(Z_{33} - \frac{\omega^2 M^2\cos^2\alpha}{Z_{22}} - \frac{\omega^2 M^2\sin^2\alpha}{Z_{11}}\right)},$$

The condition of the reading of bearing will be equality $I_3 = 0$ or the equality of zero numerator of expression for I_3. From this condition we obtain formula (4.14) in the form:

$$\operatorname{tg} \alpha_{\text{мин}} = \frac{Z_{11}}{Z_{22}} \operatorname{tg} \theta = a \operatorname{tg} \theta.$$

bearing error Δ_1 is determined by the expression

$$\operatorname{tg} \Delta_1 = \operatorname{tg} (\alpha_{\text{мин}} - \theta) = \frac{\frac{a-1}{a+1} \sin 2\theta}{1 - \frac{a-1}{a+1} \cos 2\theta} = \frac{D_1 \sin 2\theta}{1 - D_1 \cos 2\theta}. \quad (4.16)$$

From formula (4.16) it is easy to obtain the derived previously formula (4.15). Let us determine the maximum value of error Δ_1, after equating zero derivative (4.16) in terms of θ. We find $\theta_{\text{макс}}$, with which is observed maximum error $\Delta_{1\text{макс}}$:

$$\theta_{\text{макс}} = \frac{1}{2} \arccos \left(\frac{a-1}{a+1} \right). \quad (4.17)$$

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After substituting (4.17) in (4.16), we will obtain expression for the maximum error:

$$\operatorname{tg} \Delta_{1\text{макс}} = \frac{D_1}{\sqrt{1 - D_1^2}} = \frac{a-1}{2\sqrt{a}}. \quad (4.18)$$

At the value of coefficient of a, near to unity, formula (4.16) is simplified;

$$\Delta_1 \approx \frac{a-1}{a+1} \sin 2\theta \approx \frac{1}{2} (a-1) \sin 2\theta. \quad (4.19)$$

i.e. error has quadratic law with maximum values at $\theta = 45, 135, 225, 315°$. With the large dissimilarities of the amplitudes of currents within the framework (large σ) the law of a change of the error is obtained more complex: besides quadratic component appears the octantal component of error.

If we restrict error by value 0.5°, then the permissible dissimilarity of the amplitudes of the currents of field coils will be ±2o/o. It is necessary to note that in the case of $a \neq 1$ of the examination of expression for I_3 follows: resistances, introduced by frame circuits into the duct of search coil, change with the rotation of search coil of goniometer, i.e., for each position of search coil, is required its alignment of tuned circuit. This is one of the sign/criteria of the dissimilarity of the ducts of the framework with field coils.

To the same results (errors with direction finding and the staggering of search coil) gives the dissimilarity of the maximum mutual inductance of searching and the field coils of goniometer and the inequality of the effective height of the framework.

In the examined case of $a \neq 1$, $\phi = 0$, as this follows from (4.10), $A/B = 0$, i.e., the diffuseness of bearing (ellipticity of the resulting magnetic field in goniometer) are absent.

If $z_{11} = z_{22}$ or $a = 1$ and $\varphi \neq 0$, which corresponds to the case when the resistor/resistances of the ducts of the framework are differing by module, but they differ in argument, then of (4.9) and (4.12) follows that

$$\operatorname{tg} 2\alpha_{\text{мин}} = \operatorname{tg} 2\vartheta \cos\varphi$$

and

$$D = 0, \quad K = \operatorname{tg}^2 \frac{\varphi}{2} = K_2.$$

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Bearing error Δ_2 from (4.13) will be determined by the equality

$$\operatorname{tg} 2\Delta_2 = \operatorname{tg} 2(\alpha_{\text{мин}} - \vartheta) = \frac{-K_2 \sin 4\vartheta}{1 - K_2 \cos 4\vartheta} = \frac{-\operatorname{tg}^2 \frac{\varphi}{2} \sin 4\vartheta}{1 - \operatorname{tg}^2 \frac{\varphi}{2} \cos 4\vartheta}. \tag{4.20}$$

The ellipticity of field is characterized by expression (III.5):

$$\frac{A^2}{B^2} = \frac{1 - \sqrt{1 - \sin^2 2\vartheta \sin^2\varphi}}{1 + \sqrt{1 - \sin^2 2\vartheta \sin^2\varphi}}. \tag{4.21}$$

The maximum value of error $\Delta_{2\,\text{макс}}$, which occurs when

$$\vartheta_{\text{макс}} = \frac{1}{4} \arccos(-K_2),$$

it is designed from the formula

$$\mathrm{tg}\,2\Delta_{2\,\text{макс}} = \frac{K_2}{\sqrt{1-K_2^2}} = \frac{\mathrm{tg}^2\frac{\varphi}{2}}{\sqrt{1-\mathrm{tg}^4\frac{\varphi}{2}}} = \frac{1}{2}\,\frac{1-\cos\varphi}{\sqrt{\cos\varphi}}.$$

At low values ϕ, counting 1/2(1- cos ϕ) =

$$= \sin^2\frac{\varphi}{2} \approx \frac{\varphi^2}{4},\ \sqrt{\cos\varphi} \approx 1,\ \sqrt{1-\sin^2 2\theta\sin^2\varphi} \approx 1 - \frac{\varphi^2}{2}\sin^2 2\theta,$$

from (4.20) and (4.21) we will obtain

$$\Delta_2 = -1/8\varphi^2 \sin 4\theta \qquad (4.22)$$

and

$$\frac{A}{B} = 1/2\varphi \sin 2\theta \ \overset{(1)}{\text{рад}}, \qquad (4.23)$$

Key: (1). it is glad.

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At low values ϕ, the error has octantal character, with large ϕ error function is obtained more complex (4.20). If we restrict maximum error due to a phase difference by value 0.5°, then is permissible a phase difference ϕ on the order of 15°. In this case, maximum ellipticity is obtained equal to 13o/o.

Let us return to the general case when a ≠ 1 and ϕ ≠ 0. Complete bearing error can be presented in the form

$$\Delta = \alpha_{\text{мин}} - \theta = (\alpha_{\text{мин}} - \theta_1) + (\theta_1 - \theta) = \Delta_2 + \Delta_1.$$

where $\Delta_1 = \phi_1 - \theta$ - the error due to the inequality of the amplitudes of currents in field coils; $\Delta_2 = \alpha_{МНК} - \theta_1$ - error due to the presence only of phase difference ϕ.

At the low values of errors Δ_1 and Δ_2, they can be designed independently by formulas (4.19) and (4.22) respectively.

Thus, we come to the conclusion that the bearing error in the case when currents in the ducts of the framework differ in amplitude and phase and the value of error is small, equal to the sum of errors, which appear only as a result of the dissimilarity of currents in amplitude and only dissimilarity of the phases of currents. This facilitates the calculation of the error in the general case.

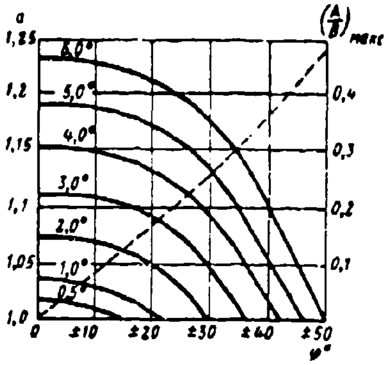

Fig. 4.15. Maximum errors of goniometric system.

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Figures by 4.15 solid lines gives to dependence of a on ϕ for the different maximum errors of goniometric system (designated above the curves). As dotted line is depicted the dependence $\left(\frac{A}{B}\right)_{макс}$ on ϕ.

2. Communication/connections between field coils of goniometer or between framework. We assume that in expression (3.44) $Z_{11} = Z_c \neq 0$. Then

$$I_2 = \frac{j\omega M E h_{дф}[\cos\theta(Z_c \cos\alpha + Z\sin\alpha) - \sin\theta(Z_c \sin\alpha + Z\cos\alpha)]}{Z_{11}(Z^2 - Z_c^2) - 2Z_c\omega^2 M^2 \cos\alpha \sin\alpha - \omega^2 M^2 Z}. \tag{4.24}$$

With $\frac{Z_c}{Z}$ real in radio direction finder, appears only the bearing error. The condition of the reading of bearing will be the equality of zero numerator (4.24), i.e.,

$$\operatorname{tg}\alpha=\frac{Z\sin\theta-Z_c\cos\theta}{Z\cos\theta-Z_c\sin\theta}=\frac{\sin\theta-\frac{Z_c}{Z}\cos\theta}{\cos\theta-\frac{Z_c}{Z}\sin\theta}. \qquad (4.25)$$

Let us find complete expression for the error function in the case when $\frac{Z_c}{Z}$ real. From (4.25) it follows that

$$\operatorname{tg}\Delta=\operatorname{tg}(\alpha-\theta)=$$

$$=\frac{\frac{Z\sin\theta-Z_c\cos\theta}{Z\cos\theta-Z_c\sin\theta}-\operatorname{tg}\theta}{1+\frac{Z\sin\theta-Z_c\cos\theta}{Z\cos\theta-Z_c\sin\theta}\operatorname{tg}\theta}$$

or

$$\operatorname{tg}\Delta=\frac{Z\sin\theta\cos\theta-Z_c\cos^2\theta-Z\sin\theta\cos\theta+Z_c\sin^2\theta}{Z\cos^2\theta-Z_c\sin\theta\cos\theta+Z\sin^2\theta-Z_c\sin\theta\cos\theta}, \qquad (4.26)$$

$$\operatorname{tg}\Delta=\frac{-\left(\frac{Z_c}{Z}\right)\cos 2\theta}{1+\left(-\frac{Z_c}{Z}\right)\sin 2\theta}.$$

Let us designate error coefficients:

And

$$-\left(\frac{Z_c}{Z}\right)=\mathrm{E} \quad \text{И} \quad \frac{1}{2}\left(\frac{Z_c}{Z}\right)^2=\mathrm{K}. \qquad (4.27)$$

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For a bearing error, we will obtain

$$\operatorname{tg}\Delta \approx \Delta \approx E\cos 2\theta + K\sin 4\theta + \ldots \qquad (4.28)$$

The values of maximum error $\Delta_{\text{макс}}$ and $\theta_{\text{макс}}$, with which is observed maximum error, they are designed from the formulas

$$\left.\begin{aligned} \operatorname{tg}\Delta_{\text{макс}} &= \Delta_{\text{макс}} = \frac{\frac{Z_c}{Z}}{\sqrt{1-\left(\frac{Z_c}{Z}\right)^2}}, \\ \theta_{\text{макс}} &= \frac{1}{2}\arcsin\left(\frac{Z_c}{Z}\right). \end{aligned}\right\} \qquad (4.29)$$

When the communication/connection between the ducts of the framework is small, i.e., $\frac{Z_c}{Z}$ is small, error has the quadratic character:

$$\Delta = E\cos 2\theta, \qquad (4.30)$$

with maximums at $\theta_{\text{макс}}$ = 0, 90, 180, 270°.

Limiting error 30', it is possible to allow $\left(\frac{Z_c}{Z}\right)$ equal not more than 0.01.

It should be noted that in multiantenna goniometric radio direction finders with the number of field coils, large of two, the presence of the communication/connection between field coils escape/ensues from the construction of goniometer. But there this communication/connection is not led to bearing errors with symmetrical antenna location and the correct winding/coil of goniometric coils.

In composite sense $\frac{Z_c}{Z}$ is obtained elliptical field in goniometer.

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If we designate $Z=R+jX$, $z=\sqrt{R^2+X^2}$, $Z_c=R_c+jX_c$, $z_c=\sqrt{R_c^2+X_c^2}$, that we will obtain formulas for the orientation of the transverse of magnetic field $\alpha_{\text{мин}}$ and for the relation of the semi-axes of the ellipse of the field:

$$\operatorname{tg} 2\alpha_{\text{мин}} = \frac{2(RR_c+XX_c)-\sin 2\theta(z^2+z_c^2)}{\cos 2\theta(z_c^2-z^2)},$$

$$\frac{2\frac{A}{B}}{1-\left(\frac{A}{B}\right)^2} = \frac{2\cos 2\theta(RX_c-XR_c)}{\sqrt{\cos^2 2\theta(z_c^2-z^2)^2+[2(RR_c+XX_c)-\sin 2\theta(z^2+z_c^2)]^2}}.$$

In conclusion let us note that the value of instrument errors depends on the circuit diagram of goniometric system. During the use of a goniometric system, it is possible or to tune both framework (or both pairs of antennas) into the adopted wave (Fig. 4.16a), or to apply untuned antennas (Fig. 4.16b). In the case of the inclined system, the least nonidentity of the circuits of two field coils or their small detuning (what it is virtually difficult to avoid) cause change in value and phase of currents in them and thereby considerable errors at direction finding. From this viewpoint, is more favorable the diagram with the unadjusted framework, since in it small changes of the parameters of circuits are not caused such the sharp fluctuations of value and phase of currents.

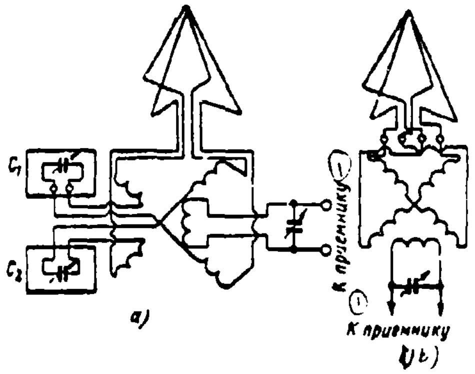

Fig. 4.16. Circuit diagrams of goniometric system: a) with spiral loops; b) with unadjusted framework.

Key: (1). To receiver.

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Therefore at present almost exceptional application/use found diagram with the unadjusted framework.

We examined the effect of breakdown of conditions (3.45), characterizing the normal operation of goniometric system, with the exception of the effect of the field nonuniformity in goniometer it is given below in the examination of the errors of goniometer.

Analysis of the effect of nonuniformity of the field in the goniometer is given below during examination of errors of the goniometer.

4.7. The errors, caused by goniometer.

The fundamental reasons for the errors of goniometer they are:

1) the dissimilarity of the magnetic fields of field coils;

2) the commutation/connection between field coils;

3) the nonperpendicularity of field coils;

4) capacitive coupling between field and search coils;

5) the nonuniformity of the field of field coils.

In § 4.6 it was establish/installed that with the dissimilarity of the magnetic fields of field coils on amplitude and on the phase in goniometer with two field coils appear the error and the ellipticity of the resulting magnetic field which are designed from formulas (4.10) and (4.11).

In goniometer with the large number of field coils, the dissimilarity of magnetic field of one of the coils causes the same errors and ellipticity independent of the number of field coils in goniometer. In goniometer with four field coils, the magnetic fields of separate coils can be distinguished between themselves to different values. Depending on the distribution of these dissimilarities, four zero directions of the pairs of coils, shifted normally one relative to another to 90°, can take any values. the permissible dissimilarities of magnetic fields in these systems approximately the same and in the simplest system with two field coils.

The effect of the communication/connection between the field coils or the framework in system with two framework is examined into § 4.6.

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The error, caused by the nonperpendicularity of the field coils of the goniometer or two framework (two pairs of the spaced antennas), is approximately expressed by the formula

$$\Delta = \frac{\delta}{2}(1 - \cos 2\theta), \qquad (4.31)$$

where δ is an angle of deflection from perpendicularity.

This formula is not considered mutual inductance between framework or coils of goniometer (with their nonperpendicularity) whose effect is examined into § 4.5.

The errors, caused by capacitive coupling between field and search coils, depend on frequency. Therefore in direction finder it is necessary to avoid the appearance of such communication/connections. For a decrease in the capacitance/capacity between field and search coils in goniometer, is applied the electrostatic shield. The application/use of a screen very substantially decreases also the value of antenna effect. The construction of goniometer with the electrostatic shield is shown in Fig. 4.17.

It was previously assumed that the field, in which rotates search coil of goniometer, is evenly.

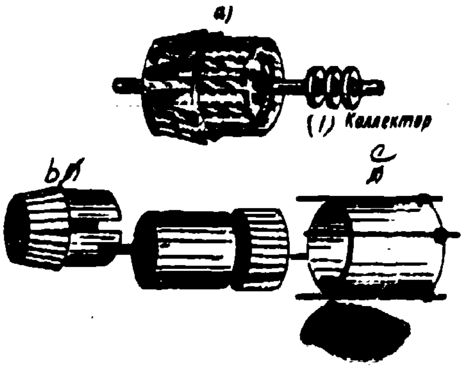

Fig. 4.17. Construction of goniometer with electrostatic shield: a) assembled selector; b) screen and framework of search coil; c) framework of field coils.

Key: (1). Collector/receptacle.

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With the nonuniformity of the field of field coils, the form of directional characteristic is not changed, but the orientation of search coil by which is obtained zero reception/procedure, it no longer corresponds accurately to the angle, composed by incoming electromagnetic field with plane of one of the framework.

From the considerations of symmetry, it is clear that in goniometer with two field coils this difference will not manifest itself, when ray/beam falls in plane of one of the framework or at an angle of 45° to them. The error function of goniometer is depicted on Fig. 4.18. This error he is called the octant error of communication/connection.

To decrease it is possible by means of the appropriate selection of geometric dimensions and form of field and search coils.

The octant error will be equal to zero with any form of field, if search coil is carried out in the form of the winding, evenly distributed in the body surface of rotation [1.6]. One form of this winding is described into § 3.9 [4.6].

The octant error can be eliminated also, if search coil is fulfilled of two sections, wound at an angle of 45° to each other. Since during rotation through 45° octant error reverses the sign and has approximately the same absolute value, the errors of two parts of this selector average out. The virtually curve of the octant error not always is is completely symmetrical, and the angle of shift of the sections of selector is necessary to select experimentally.

Figures 4.19. gives the dependence of the error of goniometer on the angle of bearing in the range of frequencies 30-100 MHz. Each field coil of goniometer consists of two turns, wound at the different angles of relatively each other. Errors are given for the angles between turns 27-90°, designated in figure.

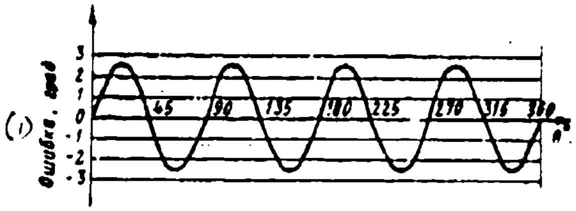

Fig. 4.18. Octant error in goniometer.

Key: (1). Error deg.

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The errors of goniometer it can be decreased by the subdivision of field or search coils on section, moreover turn number in each section makes different. The distribution of turns on sections is selected experimentally [4.4]. Figures 4.20 depicts the form of the subdivided stator.

Of goniometer with four field coils, the octant error, caused by the nonuniformity of the fields of coils, is close to zero, since is obtained not compensation. Remanent/residual error has usually duodecimal character.

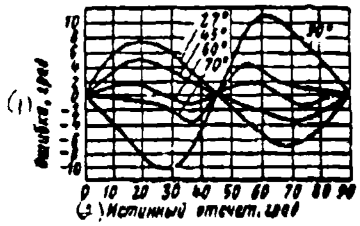

Fig. 4.19. Errors of goniometer.

Key: (1). Errors deg. (2). True reading, deg.

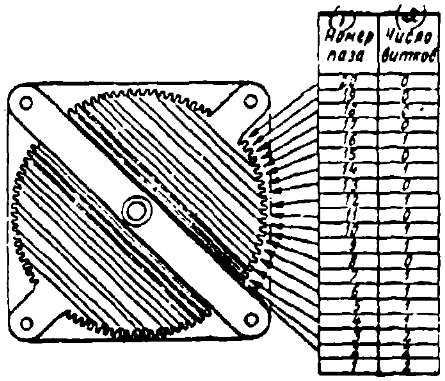

Fig. 4.20. Arrangement/permutation of turns in groove/slots of stator of goniometer.

Key: (1). Number of groove/slot. (2). Turn number.

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The errors, connected with the field nonuniformity in goniometer, are caused by the facts that in the virtually fulfilled goniometers the law of a change of the mutual inductance between field and search coils of goniometer depending on the angle between coils can differ from the sinusoidal and be periodic function. Usually more powerfully are developed the odd harmonics of communication/connection.

During the development of goniometer (or the matching coordinate transformer) for multimast radio direction finders with the auditory or automatic reading of bearing it is necessary to consider that the harmonics of the communication/connection of the coils of goniometer are led to the different errors depending on the method of the reading of bearing, number of antennas, separation of antennas and law of communication/connection. These questions are investigated in [4.1].

With use of goniometer in by antenna to system by a small separation with an increase in the number of pairs of antennas a increasing quantity of odd harmonics of communication/connection ceases to affect the creation of bearing errors. For antennas with large separation, besides the errors of separation, examine/considered into § 4.9, and the errors of goniometer, appear the supplementary errors and diffuseness in the reading of the minimum for audition (or the ellipticity of the image of bearing in the two-channel radio direction finder), that depend on the combined effect of the harmonics of communication/connection, separation of antennas and their number.

4.8. Instrument errors of system with the spaced antennas.

To system with the spaced antennas are characteristic the instrument errors, examined earlier. Direct reception/procedure, reradiation of receiver, and also the mechanical failures (slope/inclination of the axis of rotary system, eccentricity of the scale of rotary system and goniometer, incorrect engraving and inaccurate installation of the scale of the reading of bearing, etc.) they are developed in this system, just as in system with the revolving framework (see § 4.4).

Depending on that, as is utilized this system (in the form of the rotary pair of the spaced antennas or with goniometer), antenna effect has different character (see §§ 4.2 and 4.5). During the use of this system with goniometer, can appear the errors, characteristic to any goniometric system (see §§ 4.6 and 4.7).

Besides these the system with the spaced antennas possesses still other specific sources of errors.

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Important value has the error, which appears as a result of the large distance between the vertical wire antennas in motionless external system. However, for the purpose of an increase in the sensitivity, they frequently increase the separation of antennas and thereby they allow/assume this error (see § 4.9).

Let us examine the rotating couple of the spaced antennas A and B, which are located at a distance l_{r1} and l_{r2} from the receptor, arrange/located at point O. (Fig. 421).

Assume that a pair of antennas is influenced by a vertically polarized field. As the beginning for the counting off of phases we select the phase of the field at point O. Let us designate the phase of field at point A by

$$\Phi_A = -\frac{2\pi}{\lambda} l_{r1} \cos \theta,$$

the phase of field at point B by

$$\Phi_B = -\frac{2\pi}{\lambda} l_{r2} \cos \theta.$$

where θ is an angle between the direction of the oriented radio station and the plane of the pair of antennas.

The same phases Φ_A and Φ_B will have emf E_A and E_B in antennas A and B.

Let the effective height of antenna A be equal to h_A, antenna B h_B. Then

$$\dot{E}_A = \mathrm{E} h_A e^{j\Phi_A},$$
$$\dot{E}_B = \mathrm{E} h_B e^{j\Phi_B}. \qquad (4.32)$$

Currents I_A and I_B will be out of phase of relatively corresponding emf.

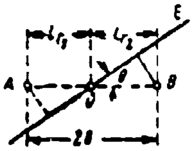

Fig. 4.21. Asymmetric location of spaced antennas.

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Phase displacement of currents, if we examine currents of field coil, i.e., of point O, will consist of two terms:

- phase displacement because of the phase angle of resistor/resistances Z_A and Z_B. the value of this phase displacement by designation ν_A and ν_B respectively:

$$Z_A = z_A e^{-j\nu_A},\quad Z_B = z_B e^{-j\nu_B}; \qquad (4.33)$$

- phase displacement ψ_A and ψ_B because of propagation time of current along the feeder, which connects antennas with the receiver:

$$\psi_A = -\frac{2\pi}{\lambda}\sqrt{\varepsilon_1}\, l_{ф1},\quad \psi_B = -\frac{2\pi}{\lambda}\sqrt{\varepsilon_2}\, l_{ф2}, \qquad (4.34)$$

where $\varepsilon_1, \varepsilon_2, l_{ф1}, l_{ф2}$ are dielectric constants and the length of feeders.

Currents are expressed:

$$\left.\begin{aligned}I_A&=\frac{\dot{E}_A}{Z_A}e^{j\psi_A}=\frac{Eh_A}{z_A}e^{j(\Phi_A+\psi_A+\nu_A)}=I_Ae^{j\varphi_A},\\ I_B&=\frac{\dot{E}_B}{Z_B}e^{j\psi_B}=\frac{Eh_B}{z_B}e^{j(\Phi_B+\psi_B+\nu_B)}=I_Be^{j\varphi_B},\end{aligned}\right\}\quad(4.35)$$

where

$$\varphi_A=\nu_A-\frac{2\pi}{\lambda}(\sqrt{\varepsilon_1}\,l_{\Phi 1}+l_{r1}\cos\theta);$$

$$\varphi_B=\nu_B-\frac{2\pi}{\lambda}(\sqrt{\varepsilon_2}\,l_{\Phi 2}-l_{r2}\cos\theta).\qquad(4.36)$$

Spill current, which causes reception/procedure, is expressed as follows:

$$I_0=(I_B\cos\varphi_B-I_A\cos\varphi_A)+j(I_B\sin\varphi_B-I_A\sin\varphi_A).\quad(4.37)$$

where the amplitude of the current

$$I_0=\sqrt{I_A^2+I_B^2-2I_AI_B\cos(\varphi_B-\varphi_A)}.$$

Let us examine several special cases.

1. Inequality of distances of antennas from receiver. In this case

$$l_{r1}\neq l_{r2},\ l_{\Phi 1}\neq l_{\Phi 2},$$

but the electrical symmetry of two antennas is retained, i.e.,

$$Z_A=Z_B=Z,\ \nu_A=\nu_B=\nu,\ h_A=h_B=h.$$

The difference current from (4.37) will be

$$I_0=\frac{Eh}{z}[(\cos\varphi_B-\cos\varphi_A)+j(\sin\varphi_B-\sin\varphi_A)]$$

or

$$I_0=-2\frac{Eh}{z}\Big[\sin\left(\frac{\varphi_B+\varphi_A}{2}\right)\sin\left(\frac{\varphi_B-\varphi_A}{2}\right)-$$
$$-j\sin\left(\frac{\varphi_B-\varphi_A}{2}\right)\cos\left(\frac{\varphi_B+\varphi_A}{2}\right)\Big]. \qquad (4.38)$$

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The amplitude of current is expressed:

$$I_0=2\frac{Eh}{z}\sin\left(\frac{\varphi_B-\varphi_A}{2}\right). \qquad (4.38')$$

Let us substitute values φ_A and φ_B by formula (4.35) into formula (4.38'):

$$I_0=2\frac{Eh}{z}\sin\frac{\pi}{\lambda}\left[(\sqrt{\varepsilon_1}\,l_{\psi 1}-\sqrt{\varepsilon_2}\,l_{\psi 2})+(l_{r1}+l_{r2})\cos\theta\right]. \qquad (4.39)$$

Current is equal to zero, when argument with sine is equal to zero, i.e., when

$$\cos\theta=\frac{\sqrt{\varepsilon_2}\,l_{\psi 2}-\sqrt{\varepsilon_1}\,l_{\psi 1}}{l_{r1}+l_{r2}}.$$

Since right side is not equal to zero, are two solutions: θ and $180°-\theta$ which are distinguished to the angle, not equal to 180°; here occurs the fracture of the axis of bearing.

Thus, we see that the asymmetric antenna location relative to receiver leads to the fracture of the axis of bearing and is developed or analogous with the phase component antenna effect (§ 4.2). The value of fracture is determined by the relative displacement of antennas from symmetrical location.

Let us determine the permissible amount of the displacement of antennas from the center, accepting the standard deviation of bearing from the true of 0.25-0.5°.

Let us assume $\theta = \theta_0 + \Delta$. In the absence of error, the current is turned into zero, when $\theta_0 = 90°$. Error $\Delta = \theta - 90°$ is determined from

$$\sin\Delta \approx \Delta = \frac{\sqrt{\epsilon_1}\, l_{\phi 1} - \sqrt{\epsilon_2}\, l_{\phi 2}}{l_{r1} + l_{r2}} = 0{,}005 + 0{,}01.$$

Let

$$\epsilon_1 = \epsilon_2 = 1,\ l_{\phi 1} - l_{\phi 2} = \Delta l_{\phi} = l_{r1} - l_{r2} = \Delta l_{r};\ l_{r1} + l_{r2} = 2b.$$

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In this case must be fulfilled the condition

$$\frac{\Delta l_{\phi}}{2b} = \frac{\Delta l_{r}}{2b} = 0{,}005 \div 0{,}01,$$

i.e. the central location of receiver must be maintained with accuracy 0.5-1o/o.

2. Asymmetry of antennas on phase angle. We set/assume:

$$\nu_A - \nu_B = \nu,$$

$$l_{\text{ф}1} = l_{\text{ф}2} = l_{\text{ф}};\ \varepsilon_1 = \varepsilon_2 = \varepsilon,\ l_{\text{г}1} = l_{\text{г}2} = b,$$

$$\sqrt{\varepsilon_1}\, l_{\text{ф}1} = \sqrt{\varepsilon_2}\, l_{\text{ф}2} = \sqrt{\varepsilon}\, l_{\text{ф}},$$

$$I_A = I_B = I = \frac{Eh}{z},$$

$$\varphi_A = \nu_A - \frac{2\pi}{\lambda}(\sqrt{\varepsilon}\, l_{\text{ф}} + b\cos\theta),$$

$$\varphi_B = \nu_B - \frac{2\pi}{\lambda}(\sqrt{\varepsilon}\, l_{\text{ф}} - b\cos\theta).$$

After substituting values φ_A and φ_B in (4.38'), we will obtain

$$I_0 = 2\frac{Eh}{z}\sin\left(\frac{2\pi}{\lambda}b\cos\theta - \frac{\nu}{2}\right). \qquad (4.40)$$

As in the preceding case, here is obtained the fracture of the axis of bearing. Error in the determination of bearing is determined by the formula

$$\sin\Delta \approx \Delta = \frac{\nu\lambda}{4\pi b}.$$

With 2b/λ = 0.2, limiting error by value Δ = 0.28°, we will obtain that a difference in the phase angles impedances of both antennas must not exceed value

$$\nu < \frac{4\pi b}{\lambda}\sin\Delta = 2\pi\cdot 0{,}2\cdot 0{,}005 = \frac{2\pi}{1000},\ \overset{(1)}{\text{или}}\ \nu < 0{,}36^\circ.$$

Key: (1). or.

From this example it is evident that the symmetry of antennas must be fulfilled with very high accuracy.

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Figures 4.21 and formula (4.37) for spill current shows that in the examined cases the different component of spill current I_0 is approximately equal to zero and reception/procedure is determined that mainly, by the nonphase component whose amplitude value is expressed by formula (4.39) or (4.40), moreover occurs the fracture of the axis of bearings.

For the limitation of the fracture of axis, it is necessary that with possible asymmetry of antennas in resistor/resistance a change in the angle v in the phase response of the resistor/resistance of antenna feeder system would be little.

It is known that especially strongly changes the phase response of any circuit at frequencies, close to resonance. It is necessary so to select the parameters of the antenna feeder system of the pair of antennas so that the frequencies of resonance for the nonphase component of current would be located outside working frequency band.

This requirement is placed during the design of antenna systems with the spaced antennas.

3. Inequality of amplitudes of currents in antennas. We assume that

$$\left.\begin{aligned} I_A = I + \Delta I;\ I_Б = I - \Delta I, \\ l_{Г1} = l_{Г2} = \frac{b}{2},\ \sqrt{\varepsilon_1}\, l_{Ф1} = \sqrt{\varepsilon_2}\, l_{Ф2},\ v_A = v_Б = v_0. \end{aligned}\right\} \quad (4.41)$$

Under these conditions

$$\varphi_Б - \varphi_A = \frac{4\pi b}{\lambda} \cos \theta.$$

After substituting values (4.41) into formula (4.37) for spill current I_0, we will obtain:

$$\begin{aligned} I_0 = -2\Big[I\Big(\sin\frac{\varphi_Б+\varphi_A}{2}\sin\frac{\varphi_Б-\varphi_A}{2}\Big) + \\ + \Delta I\Big(\cos\frac{\varphi_Б+\varphi_A}{2}\cos\frac{\varphi_Б-\varphi_A}{2}\Big)\Big] + \\ + j2\Big[I\Big(\cos\frac{\varphi_Б+\varphi_A}{2}\sin\frac{\varphi_Б-\varphi_A}{2}\Big) - \\ - \Delta I\Big(\sin\frac{\varphi_Б+\varphi_A}{2}\cos\frac{\varphi_Б-\varphi_A}{2}\Big)\Big]. \end{aligned}$$

For the amplitude of spill current, we have

$$I_0 = 2\sqrt{I^2 \sin^2\frac{\varphi_Б-\varphi_A}{2} + (\Delta I)^2 \cos^2\frac{\varphi_Б-\varphi_A}{2}}.$$

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Considering that

$$\frac{\varphi_Б-\varphi_A}{2} = \frac{2\pi b}{\lambda}\cos\theta$$

- low value, it is possible to assume:

$$\sin\frac{\varphi_B-\varphi_A}{2}=\frac{\varphi_B-\varphi_A}{2}=\frac{2\pi b}{\lambda}\cos\theta,$$

$$\cos\frac{\varphi_B-\varphi_A}{2}=1,$$

$$I_0=2\sqrt{\left(I\frac{2\pi b}{\lambda}\cos\theta\right)^2+(\Delta I)^2}. \qquad (4.42)$$

As can be seen from last/latter expression, the inequality of the amplitudes of currents in antennas on its action is analogous with the antenna effect, which has phase displacement 90°, and it leads to the formation/education of the diffuse minimums without a change in their direction.

For the limitation of value ΔI, it is expedient to select the parameters of the antenna feeder system of the pair of antennas so that the frequencies of resonance for the phase component of current would not be located in operating range. If for the purpose of sensitization in the long-wave part of the range it is necessary to disrupt this condition and to allow/assume resonance in operating range, then it is necessary to take measures for that, in order to at the resonance frequency when reactance is equal to zero, would remain sufficiently high effective resistance. For this, for example, connect of antennas supplementary effective resistance. Then the possible dissimilarities of antenna resistances do not cause the high values ΔI and large antenna effects (or errors) at direction finding.

The dissimilarity of radiation currents can occur also due to the dissimilarity of the effective height of antennas, dissimilarity of the load impedances or parameters of feeders.

We examined effect ΔI with the pair of rotary antennas; this effect in by goniometric antenna to system is examined into § 4.5.

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4.9. Error of the separation of goniometric system from 1 of the spaced antennas.

In 3 was examined the reception/procedure to the goniometric system, consisting of any number n of the spaced antennas. Is obtained formula (3.63') for the calculation of voltage/stress, induced in search coil of goniometer E_n. Current in search coil $I_n = \frac{E_n}{Z_n}$, where Z_n are resistor/resistances of the circuit of search coil.

By formula (3.63') for a current in search coil it is possible to write:

$$I_{н}=\left\{\left[\sum_{k=0}^{\infty}A_{kn+1}\sin(kn+1)\theta-\sum_{k=1}^{\infty}A_{kn-1}\sin(kn-1)\theta\right]\cos\alpha-\right.$$

$$\left.-\left[\sum_{k=0}^{\infty}A_{kn+1}\cos(kn+1)\theta+\sum_{k=1}^{\infty}A_{kn-1}\cos(kn-1)\theta\right]\sin\alpha\right\}, \tag{4.43}$$

where

$$A_{kn\pm1}=2j^{kn\pm1}J_{kn\pm1}\left(\frac{2\pi b}{\lambda}\cos\beta\right).$$

Let us present $I_{н}$ in the form [1]

$$I_{н}=M_s\cos\alpha-M_1\sin\alpha,$$

where

$$M_s=A_1\sin\theta-A_{n-1}\sin(n-1)\theta+A_{n+1}\sin(n+1)\theta-$$
$$-A_{2n-1}\sin(2n-1)\theta+A_{2n+1}\sin(2n+1)\theta+\ldots,$$
$$M_1=A_1\cos\theta+A_{n-1}\cos(n-1)\theta+A_{n+1}\cos(n+1)\theta+$$
$$+A_{2n-1}\cos(2n-1)\theta+A_{2n+1}\cos(2n+1)\theta+\ldots \tag{4.44}$$

FOOTNOTE [1]. This same to dependence is subordinated the resulting magnetic field in goniometer, and also image on the cathode-ray tube of two-channel radio range finder (see Chapter 8). ENDFOOTNOTE.

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When ratio $2\pi b/\lambda \ll 1$, then in each of the expansions M_1 and M_2 it is possible to be restricted by one member. In this case

$$I_{\text{н}} \cong J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right)(\sin\theta\cos\alpha - \cos\theta\sin\alpha) = \\ = J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right)\sin(\theta - \alpha). \qquad (4.45)$$

Current $I_{\text{н}}=0$, i.e. with auditory method is counted off bearing, when $\theta = \alpha$. Thus, is satisfied the condition of the error-free operation of the goniometric system of radio direction finder.

With an increase in the separation, equality (4.45) is disrupted. Appear the bearing error (the so-called "error of separation") and the diffuseness of reading (elliptical resulting magnetic field in goniometer).

For the calculation of the error of separation Δ and of the ellipticity of field A/B, let us turn to formula (III.12) and let us replace in it M_1 and M_2 with expressions (4.44):

$$\operatorname{tg}\left(\Delta + j\frac{A}{B}\right) = \frac{M_2\cos\theta - M_1\sin\theta}{M_1\cos\theta + M_2\sin\theta} =$$

$$= \frac{(-A_{n-1} + A_{n+1})\sin n\theta + (-A_{2n-1} + A_{2n+1})\sin 2n\theta + \ldots}{A_1 + (A_{n-1} + A_{n+1})\cos n\theta + (A_{2n-1} + A_{2n+1})\cos 2n\theta + \ldots}.$$

Assuming

$$A_1 \gg |\pm A_{n-1} + A_{n+1}| \gg |\pm A_{2n-1} + A_{2n+1}| \ldots,$$

we obtain, that

$$\operatorname{tg}\left(\Delta + j\frac{A}{B}\right) = \frac{-A_{n-1} + A_{n+1}}{A_1}\sin n\theta -$$

$$- \frac{-A_{n-1} + A_{n+1}}{A_1}\sin n\theta\,\frac{A_{n-1} + A_{n+1}}{A_1}\cos n\theta +$$

$$+ \frac{-A_{2n-1} + A_{2n+1}}{A_1}\sin 2n\theta + \ldots$$

or

$$\operatorname{tg}\left(\Delta + j\frac{A}{B}\right) = \frac{A_{n+1} - A_{n-1}}{A_1}\sin n\theta + \left(\frac{A_{n-1}^2 - A_{n+1}^2}{2A_1^2} + \right.$$

$$\left. + \frac{A_{2n+1} - A_{2n-1}}{A_1}\right)\sin 2n\theta + \ldots \qquad (4.46)$$

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Let us examine two individual cases: the number of antennas n even and the number of antennas n odd.

1. Number of antennas and even. In this case the right side of expression (4.46) is purely real; therefore A/B = 0; i.e. the ellipticity of field is absent.

Replacing A_1, $A_{n\pm1}$ and $A_{2n\pm1}\ldots$ by their expressions with Bessel functions (3.51), we obtain

$$\operatorname{tg}\Delta=\frac{(-1)^{\frac{n}{2}}\left[J_{n+1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)+J_{n-1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)\right]}{J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right)}\sin n\theta+$$
$$+\left\{\frac{\left[J_{n-1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)\right]^2-\left[J_{n+1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)\right]^2}{2\left[J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right)\right]^2}+\right.$$
$$\left.+\frac{J_{2n+1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)+J_{2n-1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)}{J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right)}\right\}\sin 2n\theta+\ldots \tag{4.47}$$

Approximately

$$\operatorname{tg}\Delta\approx\frac{(-1)^{\frac{n}{2}}\left[J_{n+1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)+J_{n-1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)\right]}{J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right)}\sin n\theta. \tag{4.48}$$

It is known that

$$J_{n-1}(a)+J_{n+1}(a)=\frac{2n}{a}J_n(a).$$

Utilizing this equality, we obtain from (4.48):

$$\operatorname{tg}\Delta \approx \frac{2n(-1)^{\frac{n}{2}} J_n\left(\frac{2\pi b}{\lambda}\cos\beta\right)}{\frac{2\pi b}{\lambda}\cos\beta\, J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right)}\sin n\theta. \qquad (4.49)$$

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Thus, we see that with even number of antennas, if is not satisfied for the separation of antennas condition $2\pi b/\lambda \ll 1$, occurs the bearing error, which is expressed by formula (4.49). The law governing the error of the separation of the antennas of dual order relative to the number of antennas, i.e., at by the four-antenna of system, these are the octant error, at by the six-antenna of system - the error of 12-fold order, etc.

During parallel connection of the pairs of adjacent antennas, which are located at an angle 2γ, from formula (4.48) we will obtain

$$\operatorname{tg}\Delta = \frac{(-i)^{\frac{n}{2}}\left[J_{\frac{n}{2}-1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)\cos\left(\frac{n}{2}-1\right)\gamma + J_{\frac{n}{2}+1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)\cos\left(\frac{n}{2}+1\right)\gamma\right]}{J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right)\times\cos\gamma}\sin\frac{n}{2}\theta. \qquad (4.50)$$

Expression (4.50) is obtained under the assumption that impedances of all antennas are identical. In reality, with asymmetric antenna location their impedances become different. The procedure of calculation of the error of separation taking into account mutual impedances with any antenna location is given in [4.9].

Formula (4.49) for the low values of the separation of the antennas when $2\pi b/\lambda \cos\beta < 1$ and $J_p(x) \approx \frac{x^p}{p!2^p}$, assumes the form

$$\operatorname{tg}\Delta \approx \frac{\left(\frac{2\pi b}{\lambda}\cos\beta\right)^{(n-2)}}{(n-1)!\,2^{(n-2)}}\sin n\theta. \qquad (4.51)$$

From formula (4.51) follows that for by the four-antenna of system, i.e., when n = 4, and with $2\pi b/\lambda \cos\beta < 1$

$$\operatorname{tg}\Delta \approx \frac{1}{24}\left(\frac{2\pi b}{\lambda}\cos\beta\right)^2\sin 4\theta. \qquad (4.52)$$

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The error of the separation at by the four-antenna of system has octant law from 9 and is designed from formula (4.52).

Figures 4.22 depicts the errors of separation depending on $S = 2\pi b/\lambda$ for the different number $n(\beta = 0)$, constructed according to formula (4.49).

On the axis of abscissas besides scale S, is plotted/applied another scale $2b/\lambda$.

Curves in Fig. 4.22 show that the errors rapidly decrease with an increase in the number of antennas and with respect rapidly increases the permissible separation for the assigned maximum error of separation.

For the separation, close to $2b = 1.22\lambda$, when $J_1(2\pi b/\lambda) = 0$ and effectiveness of system equal to zero, errors strongly grow/rise, but they rapidly fall with a decrease in the separation ($2b < 1.22\lambda$). For separation $2b = 1.8\lambda/\pi = 0.573\lambda$, the values for $J_1(2\pi b/\lambda)$ and the effectiveness of system reach maximum values.

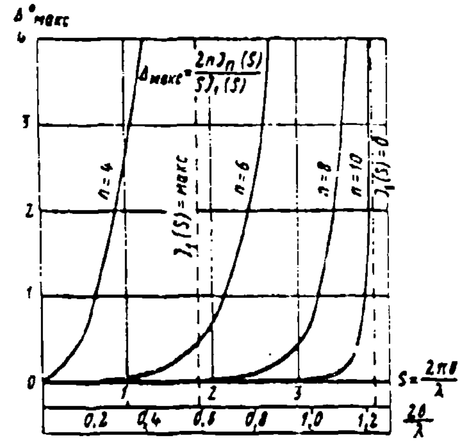

Fig. 4.22. Errors of separation with different number of antennas.

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Figures 4.23 shows close to the maximum of the error for θ = 22°30' for 8-antenna system with parallel connection of the adjacent pairs of antennas (γ = 22°30'; 27°30'; 30°) in the case when, the resistor/resistances of all antennas are considered identical, and when γ = 27°15' for one special case of antenna location, when is considered the dissimilarity of these resistor/resistances [4.9].

Here for a comparison are given errors for by the four-antenna of system. From Fig. 4.23 it follows that if we do not consider the dissimilarity of complete antenna resistances, then between the pairs of the parallel-connected antennas optimum angle will be γ = 27°30'. When the resistor/resistance of all antennas it is not possible to consider identical, the best angle between the parallel-connected antennas depends on size/dimensions by the antenna of system and parameters of single antennas. Optimum angle γ can differ from 27°30' - it must be establish/installed as a result of calculations [4.9].

To apply it is more than 8 antennas virtually inexpediently, since, limiting maximum error by value 2°, with 8 antennas it is possible to allow separation $2b = 1.05\lambda$, close to maximum permissible.

All the curves of Fig. 4.22 and 4.23 are designed for the ground wave when $\beta = 0$. Of system without parallel connection of antennas with an increase in the angle β, the error of separation falls, since in this case seemingly decreased equivalent separation.

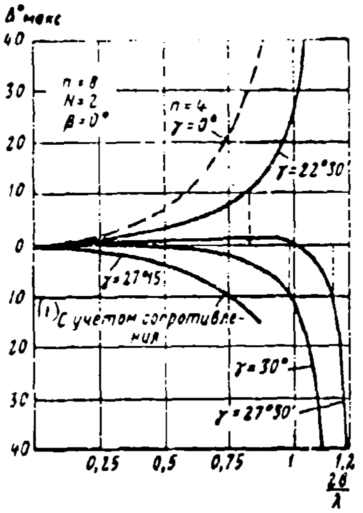

Fig. 4.23. Errors of separation of 8-mast system during parallel connection of adjacent antennas.

Key: (1). Taking into account resistor/resistance.

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In systems with the parallel-connected pairs of antennas, the error of separation not always does fall with an increase in the angle of incidence. It can have a maximum with $\beta \neq 0$, and with certain β change sign. This is explained to the facts that in such systems by selection γ is realized error compensation of separation only for the angle of the slope of a front of wave $\beta = 0$.

Table 4.1 gives the permissible separations of antennas at the different values of maximum errors for the different number of antennas n. For 8-antenna system with the parallel-connected antennas, it is taken into account, that at some angles of incidence the error can change its sign, and therefore the permissible separation for these systems is designed on the basis of the maximum spread/scope of error. Antenna resistances are assumed to be identical.

From all that has been previously stated, it follows that with an increase in the number of antennas with just one separation of antennas the error of separation falls. If we restrict error by the determined value, then the permissible separation increases with an increase in the number of antennas.

Table 4.1 gives the ratio of maximum separation of system with n antennas to the maximum separation at by the four-antenna of system.

Separation 2b = 1.22λ, with which the error of separation Δ independent of the number of antennas approaches very large value and effectiveness is equal to zero, is maximum. With 2b > 1.22λ the error changes sign and it decreases.

The ratio of the maximum separation of antennas for n-antenna system to the maximum separation of antennas at by the four-antenna of system falls with an increase in the permissible error of separation.

During the use of parallel connection of adjacent antennas for 8-antenna system, the smallest error of separation is obtained, when the angle between the parallel-connected adjacent antennas $\gamma = 27.5°$, if we do not consider the dissimilarity of antenna resistances; if we consider this dissimilarity, then $\gamma_{опт}$ depends on the parameters of concrete/specific/actual system.

2. Number of antennas n (odd. With the odd number of antennas, the right side of expression (4.46) has real and imaginary parts. The first determines error, the second - diffuseness of bearing is equal to the ratio of the axes of the ellipse of the field of goniometer.

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Table 4.1. Characteristics of goniometric type multimast antenna systems.

(1) Число антенн *n*	(2) Число полевых катушек гониометра *N*	(3) Половина угла между парой антенн γ°	(4) Номер рисунка	(5) Отношение максимально допустимого разноса к длине волны $\frac{2b}{\lambda}$ при максимальной ошибке или максимальном размахе ошибок, град				(6) Отношение максимальных разносов у *n* системы и у четырехантенной системы при $\Delta_{макс}$, град			
				0,5	1	2	10	0,5	1	2	10
4	2	0	4,22	0,15	0,2	0,28	0,57	1,0	1,0	1,0	1,0
6	3	0	4,22	0,6	0,66	0,75	1,0	4,0	3,3	2,7	1,76
8	4	0	4,22	0,97	1,0	1,05	1,15	6,4	5	3,75	1,96
10	5	0	4,22	1,16	1,2	1,21	1,22	7,8	6	4,3	2,14
8	2	22,5	4,23			0,43				1,54	
8	2	27,5	4,23			1,04	1,14			3,73	2
8	2	28,5	4,23			0,9				3,23	
8	2	30	4,23			0,75				2,68	

Key: (1). Number of antennas n. (2). Number of field coils of goniometer N. (3). One-half angle between the pair of antennas γ°. (4). Number of figure. (5). Ratio of the maximum permissible separation to wavelength 2b/λ with maximum error or the maximum spread/scope of errors, deg. (6). Relation of maximum separations of n of system and at by the four-antenna of system when $\Delta_{макс}$, deg.

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Error Δ_p with the odd number of antennas is small and is expressed (4.46)

$$\operatorname{tg}\Delta \approx \left(\frac{A_{n-1}^2 - A_{n+1}^2}{2A_1^2} + \frac{A_{2n+1} - A_{2n-1}}{A_1} \right) \sin 2n\theta \quad (4.53)$$

or

$$\operatorname{tg}\Delta_{макс} \approx \Delta_{макс} \approx \frac{\left[J_{n-1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)\right]^2 - \left[J_{n+1}\left(\frac{2\pi b}{\lambda}\cos\beta\right)\right]^2}{2\left[J_1\left(\frac{2\pi b}{\lambda}\cos\beta\right)\right]^2}. \quad (4.54)$$

The diffuseness of bearing, characterized by the ratio of the axes of the ellipse of field in goniometer, is designed with $\beta = 0$ by the formula

$$\frac{A}{B} = D \approx \frac{A_{n+1} - A_{n-1}}{jA_1} \sin n\theta \approx \frac{2nJ_n\left(\frac{2\pi b}{\lambda}\right)}{\frac{2\pi b}{\lambda} J_1\left(\frac{2\pi b}{\lambda}\right)} \sin n\theta. \quad (4.55)$$

The dominant role with the odd number of antennas plays the diffuseness of bearing. It is determined from the same formulas, as the error in the case even number of antennas (4.49) and (4.55).

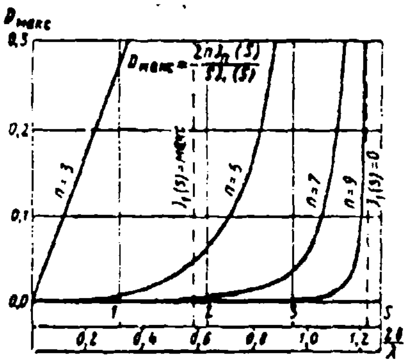

Fig. 4.24. Antenna system with the odd number of antennas. Dependence $D_{макс}$ on $S = 2\pi b/\lambda$.

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On Fig. 4.24, is depicted the dependence $D_{макс}$ on separation with the different (by odd number) number of antennas. Along the axis of abscissas, is deposit/postponed by $S = 2\pi b/\lambda$, and also the ratio $2b/\lambda$.

The odd number of antennas is not virtually applied mainly due to the appearance of antenna effects of system. The in principle minimum number of antennas of goniometric radio direction finder is equal to three.

4.10. Mounting errors of the goniometric system of the spaced antennas.

Error in orientation of antennas.

On Fig. 4.25, is represented the arrangement/permutation of antennas by a four-antenna of system. Direction in one of the antennas (the first) is displaced in the angle δ of relatively initial reference line OO_1.

Of emf, induced in antennas with the vertical electric field E of the electromagnetic wave, which comes in from azimuth θ with the angle of the slope of a front of wave β, they will be

$$\dot{E}_1 = Eh\,e^{j\frac{2\pi h}{\lambda}\cos(\theta-\delta)\cos\beta}, \quad \dot{E}_2 = Eh\,e^{j\frac{2\pi h}{\lambda}\sin\theta\cos\beta},$$
$$\dot{E}_3 = Eh\,e^{-j\frac{2\pi h}{\lambda}\cos\theta\cos\beta} \text{ and } \dot{E}_4 = Eh\,e^{-j\frac{2\pi h}{\lambda}\sin\theta\cos\beta},$$

where h is the effective height of antenna taking into account the angle of the slope of a front of wave β.

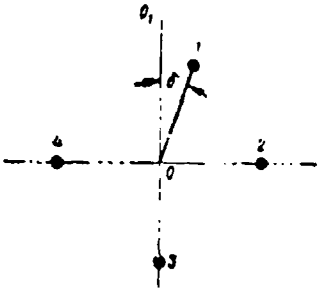

Fig. 4.25. Incorrect orientation of antenna.

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At the low value δ of emf in the first antenna, it will be

$$\dot{E}_1 = \dot{E}he^{j\left(\frac{2\pi b}{\lambda}\cos\theta\cos\beta + \frac{2\pi b\delta}{\lambda}\sin\theta\cos\beta\right)};$$

i.e. it has supplementary phase displacement $\phi = 2\pi b\delta/\lambda \sin\theta \times \cos\beta$, moreover in view of smallness δ cos φ ~ 1 and $\sin\phi = \phi = 2\pi b\delta/\lambda \sin\theta \cos\beta$. Emf in the pairs of antennas 2-4 and 1-3 in the case when $2\pi b/\lambda \ll 1$, will be

$$\dot{E}_{24} = \dot{E}_2 - \dot{E}_4 = j2\dot{E}h\frac{2\pi b}{\lambda}\cos\beta\sin\theta,$$

$$\dot{E}_{13} = \dot{E}_1 - \dot{E}_3 = j(E'_{13} + jE''_{13}),$$

where

$$E'_{13} = 2\dot{E}h\left(\frac{2\pi b}{\lambda}\cos\theta\cos\beta + \frac{\varphi}{2}\right) =$$

$$= 2\dot{E}h\frac{2\pi b}{\lambda}\left(\cos\theta + \frac{\delta}{2}\sin\theta\right)\cos\beta$$

and

$$E''_{13}=\frac{\pi b\delta}{\lambda}\sin\theta\cos\beta E'_{13}.$$

In system appear the error and the diffuseness of bearing. For small angles δ, the error is caused by value E'_{13}, diffuseness - by value E''_{13}. The calculated bearing α is determined by the equality

$$\operatorname{tg}\alpha=\frac{E_{13}}{E'_{13}}=\frac{\sin\theta}{\cos\theta+\frac{\delta}{2}\sin\theta}. \qquad (4.56)$$

Bearing error is designed from the formula

$$\operatorname{tg}|\Delta|=\operatorname{tg}|\alpha-\theta|=\frac{\frac{\delta}{2}(1-\cos 2\theta)}{1+\frac{\delta}{4}\sin 2\theta}. \qquad (4.57)$$

After equating zero derivative of expression (4.57) in terms of θ, let us find azimuth $\theta_{\text{макс}}$, when occurs maximum error $\Delta_{\text{макс}}$:

$$\theta_{\text{макс}}=90^\circ+\frac{1}{2}\arcsin\frac{\delta}{4}\approx 90^\circ$$

and

$$\Delta_{\text{макс}}=\delta.$$

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The diffuseness of the minimum is characterized by the formula

$$\frac{A}{B}=\frac{\pi b\delta}{\lambda}\sin\theta\cos\beta. \qquad (4.58)$$

Maximum value $\left(\frac{A}{B}\right)_{\text{макс}}=\frac{\pi b\delta}{\lambda}\cos\beta$ is obtained at $\theta = 90°$. From formulas (4.57) and (4.58) follows that if we restrict error $\Delta_{\text{макс}}$ by value 1°, then are the permissible value of the angle of shift of antenna $\delta \leqslant 1°$, also, in this case $\left(\frac{A}{B}\right)_{\text{макс}}=\frac{1}{57}\frac{\pi b}{\lambda}$.

Dissimilarity of the level of antenna mounting.

Let the first antenna be elevated on Δh above the remaining. Other errors in installation there are no antennas. Is oriented electromagnetic field, normally polarized.

Emf of the first antenna has supplementary phase displacement φ, caused by the facts that it is elevated:

$$\varphi=-\frac{2\pi\Delta h}{\lambda}\cos\theta\sin\beta.$$

Since $2\pi\Delta h/\lambda << 1$, then $\sin \phi/2 \approx \phi/2 = \pi\Delta h/\lambda \cos \theta \sin \beta$ and $\cos \phi/2 = 1$.

Emf in the pairs of antennas 1-3 and 2-4 will be

$$\dot{E}_{13} = j(E'_{13} + jE''_{13}), \quad \dot{E}_{24} = j2Eh\frac{2\pi b}{\lambda}\sin\theta\cos\beta,$$

moreover

$$E'_{13} = 2Eh\frac{2\pi b}{\lambda}\cos\theta\cos\beta\left(1 - \frac{\Delta h}{2b}\operatorname{tg}\beta\right)$$

and

$$E''_{13} = \frac{\pi\Delta h}{\lambda}\cos\theta\sin\beta\, E'_{13}.$$

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The angle of bearing α is determined by the equality

$$\operatorname{tg}\alpha = \frac{E_{24}}{E'_{13}} = \frac{\sin\theta}{\cos\theta\left(1 - \frac{\Delta h}{2b}\operatorname{tg}\beta\right)} \approx \left(1 + \frac{\Delta h}{2b}\operatorname{tg}\beta\right)\operatorname{tg}\theta. \quad (4.59)$$

Bearing error on the basis (4.19) can be designed by the formula

$$\Delta \approx \frac{\Delta h}{4b}\operatorname{tg}\beta\sin 2\theta \quad \text{and} \quad \Delta_{\text{макс}} = \frac{\Delta h}{4b}\operatorname{tg}\beta. \quad (4.60)$$

Diffuseness of the bearing

$$\left(\frac{A}{B}\right)_{\text{макс}} = \frac{\pi\Delta h}{\lambda}\sin\beta. \quad (4.61)$$

If the separation of antennas 2b = 10 m, Δh = 20 cm, β = 45°, then

$$\Delta_{\text{макс}} = \frac{0,2}{20} \approx 0,6^\circ.$$

When λ = 30 m $\left(\frac{A}{B}\right)_{\text{макс}} \approx 0,015.$

Dissimilarity of a radius of the arrangement/permutation of antennas.

For the pair of antennas, this case is examined into § 4.8. It howled shown, that with a difference in the distances of antennas of center Δb appears the phase antenna effect $a \approx \Delta b/b$.

The phase antenna effect, which appears of the pair of antennas, is brought in goniometric system to the bearing error whose maximum value is calculated by formula (4.47):

$$\Delta_{\text{макс}} = \frac{a}{\sqrt{1-4a^2}}.$$

So that the error will be not more than 0.5°, distances of center must be maintained with accuracy Δb/b < 0.008.

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Slope/inclination of one of the antennas.

The first antenna has slope angle toward vertical line

Slope/inclination can have any orientation relative to the plane of the pair of antennas 1-3. Let us examine two limiting cases: slope/inclination in the plane of the pair of antennas 1-3 and slope/inclination in the plane, perpendicular to the plane of the pair of antennas 1-3.

Let us determine error for the first case. Is oriented field with strength E and the angle of polarization γ. Slope of a front of wave β. The effective height of the first antenna for the reception of the vertical field

$$h_{\text{в}} \approx h\cos\beta$$

and for the reception of the horizontal field, perpendicular to the plane of propagation,

$$h_{\text{г}} \doteq h\sin\eta.$$

The amplitude of emf in a pair of antennas 1-3 will be:

from the vertical field

$$E'_{13}=\frac{4\pi bhE}{\lambda}\cos^2\beta\cos\gamma\cos\theta,$$

from the horizontal field

$$E''_{13}=Eh\sin\eta\sin\gamma\sin\theta.$$

We assume that emf E'_{13} and E''_{13} coincide in phase. Then the amplitude of complete emf in a pair 1-3 we obtain as sum E'_{13} and E''_{13}:

$$E_{13}=E'_{13}+E''_{13}=\frac{4\pi bhE}{\lambda}\cos^2\beta\cos\gamma\times$$
$$\times\left[\cos\theta+\frac{\sin\eta\operatorname{tg}\gamma}{\frac{4\pi b}{\lambda}\cos^2\beta}\sin\theta\right].$$

The amplitude of emf in a pair of antennas 2-4 will be

$$E_{24}=\frac{4\pi bhE}{\lambda}\cos^2\beta\cos\gamma\sin\theta.$$

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The reading of bearing α on radio direction finder is determined by the expression

$$\operatorname{tg}\alpha=\frac{E_{24}}{E_{13}}=\frac{\sin\theta}{\cos\theta+a\sin\theta},$$

where

$$a=\frac{\sin\eta\,\mathrm{tg}\,\gamma}{\frac{4\pi b}{\lambda}\cos^2\beta}.$$

Bearing error will be

$$\mathrm{tg}\,\Delta\approx\Delta=\mathrm{tg}\,(\alpha-\theta)=\frac{\frac{a}{2}(1-\cos 2\theta)}{1+\frac{a}{2}\sin 2\theta}. \qquad (4.62)$$

The maximum value of the error when $\theta = 90°$ is equal to

$$\Delta_{\text{макс}}=a=\frac{\sin\eta\,\mathrm{tg}\,\gamma}{\frac{4\pi b}{\lambda}\cos^2\beta}. \qquad (4.63)$$

Let $\eta=5°$, $\gamma = 45°$, $\beta = 45°$, $b/\lambda = 0.1$, then

$$\Delta_{\text{макс}}=\frac{0{,}087\cdot 1}{1{,}26\cdot 0{,}5}=0{,}138\approx 8°.$$

In the case of the slope/inclination of the first antenna in the plane, perpendicular to the plane of the pair of antennas 1-3, the condition for the reading of bearing will be

$$\mathrm{tg}\,\alpha=\frac{\sin\theta}{(1+a)\cos\theta} \qquad (4.64)$$

and bearing error $\Delta_{\text{макс}}=\frac{a}{2}$.

In other cases the maximum error is located from a/2 to a.

We assumed that the phases of emf from horizontal and vertical fields coincide. In reality between them, there can be a phase difference. Then error decreases, appears the diffuseness of the minimum.

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Examined will be requirements for installation the four-antenna of goniometric system. The conclusions about the permissible inaccuracies in the mounting of antennas approximately are retained for the case of multimast radio direction finder.

4.11. Symmetrization of diagram with the connection of antennas by coaxial cable.

In radio direction finder with motionless antenna by system are applied several antennas - the vertical vibrators or the framework.

The connection of the symmetrical directional antenna to the asymmetric input of radio direction finder, and also the connection of separate symmetrical vibrators to common asymmetric diagram is convenient to realize by single-cable coaxial cable. So that in this case is not disrupted the symmetry of the diagram of the directional antenna of vibrators (did not appear the single-cycle current), are applied conversion transformers (on long, average and short waves) and the balancing cell/elements (in the VHF range). Besides the transformers of the described systems into §4.3 it is possible application/use and other constructions [4.5, 4.10].

There is special interest in the transformer whose windings are coiled around ferrite tori. The communication/connection between windings is obtained because of currents (field) in the shielding jacket. These transformers with coupling coefficient, close to 1, provides the large symmetry of connected to winding symmetric loading [4.5].

The simplest systems of balancers are depicted on Fig. 4.26.

In diagram with U-bend (Fig. 4.26a) to the end/leads of balanced network are connected the end/leads of the supplementary cable, which forms U-bend. The vein/strand of 1 coaxial cable of feed is connected to one end/lead of the load, the sheathing of 2 cables is connected with the sheathing of U-bend and it is grounded. Because of the fact that the length of U-bend is selected as being equal to $\lambda/2$, the phases of stresses on the end/leads of balanced network differ on 180° and single-cycle current is absent. If resistance of balanced network are designated $z_н$. then for the agreement of loads the wave impedance of coaxial cable must be equally to $\frac{z_н}{4}$.

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In the construction, depicted on Fig. 4.26b, the vein/strand of coaxial cable 1 and sheathing 2 are connected to the end/leads of symmetric loading. To sheathing is placed with small gap metallic beaker (3-3) by length λ/4, opened from load and soldered to cable sheathing from another side (a-a). Beaker forms with cable sheathing the short-circuited line with entry impedance $Z_{ст}=j\rho_{ст}\,\mathrm{tg}\,\frac{2\pi}{\lambda}l$ also, at length $l=\frac{\lambda}{4}$ $Z_{ст}=\infty$. In this case, the current from the end/lead of 2 loads on cable sheathing is absent and the symmetries of diagram is not disrupted.

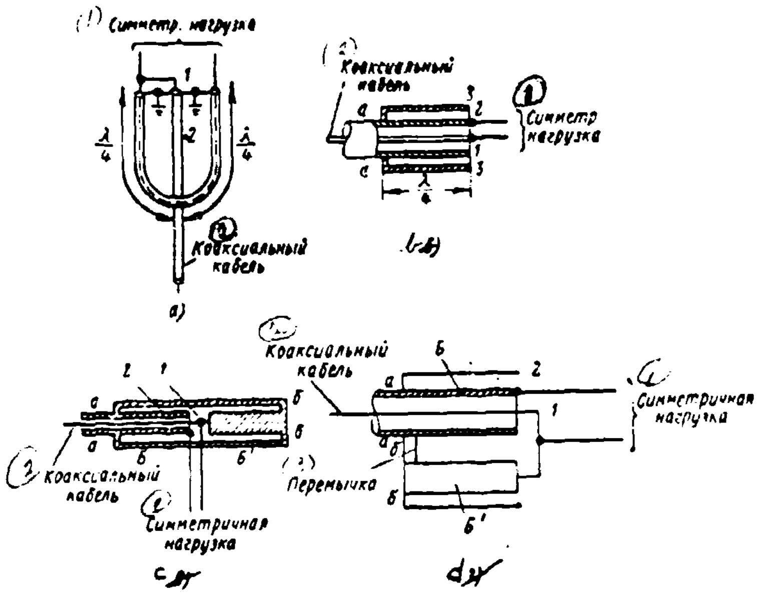

Fig. 4.26. Equivalent diagrams of goniometric system of two framework.

Key: (1). Symmetrical. (2). Coaxial cable. (3). Cross connection.

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The diagrams of U-bend and beaker are narrow-band. In the device, depicted on Fig. 4.26c, vein/strand 1 and the sheathing of 2 cables are connected to symmetric loading $Z_н$ just as on Fig. 4.26b. To vein/strand 1, is connected the even metallic cylinder whose diameter is equal to the diameter of cable sheathing. To cylinder and cable sheathing, is put on (with a small gap) the metal tube, soldered to them in a-a and b-b and generating the line, short-circuited from two sides. Since the halves of line B and B' have identical length, wires 1 and 2 prove to be equally loaded and the symmetry of diagram is not disrupted. Diagram Fig. 4.26c is wide-range and it is used in the section of range, while short-circuit impedance is greater than load impedance.

On Fig. 4.26d, is given the schematic of the modified device, shown to Fig. 4.25c; it differs in the facts that the halves of the short-circuited line B and B' for the purpose of a decrease in the overall sizes are bent. With the movement of the supplementary cross connection of short circuit it is possible to change wavelength with which back-out resistor of line is equal to infinity.

4.12. Effect of the dissimilarity of the parameters of the connecting cables.

For the correct work of radio direction finder, the cables must not change the amplitude ratio and phases of emf, the induced within the separate antennas or framework. At the same time, the pieces of cable, serving for the connection of the separate antennas and framework as the antenna of the system of radio direction finder, sometimes have the different parameters. The lengths of the pieces of cables, as thoroughly they had not selected, also can be distinguished. For this reason is disrupted the normal operation of the radio direction finder: appear the instrument errors and the diffuseness of the minimum with the reading of bearing to audition or ellipticity of image in two-channel visual receiving indicators.

The dissimilarities of wave impedance ρ_ϕ and of the attenuation β_ϕ of cables are usually small and can be disregarded. The dissimilarities of the coefficient of elongation $\xi=\sqrt{\epsilon}$, where ϵ — the equivalent dielectric constant of cable, and geometric lengths they lead to the fact that are distinguished between themselves the electrical lengths of cables.

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Let us designate the dissimilarity of the electrical lengths of the pieces of the cable through Δml, where $m=\frac{2\pi}{\lambda}\sqrt{\epsilon}$.

Effect of the dissimilarity of the electrical lengths of cables in a goniometric system.

On Fig. 4.27, is depicted the equivalent diagram of system of two framework (or two pairs of the spaced antennas). We assume that all network elements, with the exception of the lengths of cables of the framework, are identical.

Let us designate:

$$E_1 = E_0 \cos\theta \quad \text{and} \quad E_2 = E_0 \sin\theta$$

- emf, induced within the framework or pairs of opposite antennas;

Z_a — the load impedance of cable from antennas;

$Z_n = r_n + jX_n$ — the load impedance of cable from receiver;

ml and $ml + \Delta ml$ — the electrical lengths of the cables, which go from the pairs of antennas.

Let us convert by voltage E_1 and resistor/resistance Z_a toward the end of the cable where is included the load of the receiver:

$$\dot{E}'_1 = \frac{\rho_\phi E_1}{\rho_\phi \cos ml + jZ_a \sin ml}, \tag{4.65}$$

$$Z'_a = \rho_\phi \frac{Z_a \cos ml + j\rho_\phi \sin ml}{\rho_\phi \cos ml + jZ_a \sin ml}. \tag{4.66}$$

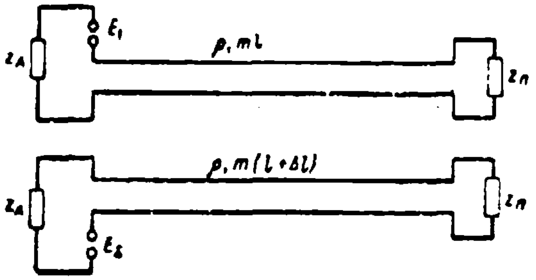

Fig. 4.27. equivalent diagram of the goniometric system of two framework.

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Current in the load of the cable of the first framework will be [1]

$$I_1 = \frac{\dot{E}'_1}{Z'_{\text{A}} + Z_{\text{н}}}. \qquad (4.67)$$

FOOTNOTE [1]. In the case of the presence in diagram, let us examine the torque/moment of the reading of bearing, when there is no reaction of search coil to the ducts of the framework. ENDFOOTNOTE.

After the substitution in (4.67) of formulas (4.65) and (4.66) we will obtain

$$\dot{I}_1 = \frac{\rho_\Phi \dot{E}_1}{\rho_\Phi (Z_n + Z_a)\cos ml + i(Z_a Z_n + \rho_\Phi^2)\sin ml}. \qquad (4.68)$$

Analogous expression can be written for current $\dot{I}_2$, on the basis of emf E_2 and length of cable $l_2 = l + \Delta l$.

Let us examine two special cases:

$$Z_a = \rho_\Phi,$$

then

$$\dot{I}_1 = \frac{E_1}{\sqrt{(R_n + \rho_\Phi)^2 + X_n^2}}\, e^{-i\left(ml + \operatorname{arctg}\frac{X_n}{R_n + \rho_\Phi}\right)},$$

and

$$Z_n = \rho_\Phi,$$

then

$$\dot{I}_1 = \frac{E_1}{\sqrt{(R_a + \rho_\Phi)^2 + X_a^2}}\, e^{-i\left(ml + \operatorname{arctg}\frac{X_a}{R_a + \rho_\Phi}\right)}.$$

In both cases, i.e., when from any side of cable, load impedance is equal to the wave impedance of cable, the amplitude of current in load does not depend on length l. The length of cable determines the phase of current. The dissimilarity of the lengths of the cables of the framework in these cases is brought only to the diffuseness of the minimum (to the ellipticity of the resulting magnetic field in goniometer).

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In the general case of resistor/resistance Z_a and Z_n they have composite character. By addition at the cable of certain length a can make equivalent resistance Z'_n (4.66) by purely active and equal to $Z'_n = S\rho_{\phi}$. From the course of antennas, it is noted [for 3.1] that

$$\operatorname{tg} 2ma = \frac{2\rho_{\phi} X_a}{R_a^2 + X_a^2 - \rho_{\phi}^2}, \qquad (4.69)$$

$$S^2 = \frac{1 - \sqrt{1 - \dfrac{4R_a^2 \rho_{\phi}^2}{(R_a^2 + X_a^2 + \rho_{\phi}^2)^2}}}{1 + \sqrt{1 + \dfrac{4R_a^2 \rho_{\phi}^2}{(R_a^2 + X_a^2 + \rho_{\phi}^2)^2}}} =$$

$$= \frac{R_a^2 + X_a^2 + \rho_{\phi}^2 - \sqrt{(R_a^2 + X_a^2 + \rho_{\phi}^2)^2 - 4R_a^2 \rho_{\phi}^2}}{R_a^2 + X_a^2 + \rho_{\phi}^2 + \sqrt{(R_a^2 + X_a^2 + \rho_{\phi}^2)^2 + 4R_a^2 \rho_{\phi}^2}}. \qquad (4.70)$$

Here S - KBV of cable, determined by antenna resistance. Let us designate the new equivalent length of cable $l_0 = l + a$. From (4.68) the expression for a current in the load of the cable of the first framework assumes the form

$$I'_1 = \frac{E_1}{(Z_n + S\rho_{\phi}) \cos ml_0 + j(\rho_{\phi} + SZ_n) \sin ml_0} = I'_{10} \cos \theta,$$

where

$$I'_{10} = \frac{E_0}{(Z_n + S\rho_{\phi}) \cos ml_0 + j(\rho_{\phi} + SZ_n) \sin ml_0} \qquad (4.71)$$

or

$$(I'_{10})^2 \approx \frac{E_0^2}{(Z_{н} + S\rho_{ф})^2 \cos^2 ml_0 + (\rho_{ф} + SZ_{н})^2 \sin^2 ml_0}. \qquad (4.71')$$

The phase of current I'_{10} in load $Z_{н}$ is determined from (4.71) by the expression

$$\varphi_1 = \operatorname{arc\,tg}\left(\frac{\rho_{ф} + SZ_{н}}{Z_{н} + S\rho_{ф}} \operatorname{tg} ml_0\right). \qquad (4.72)$$

If the lengths of the cables of both of the framework are identical, then currents $I'_{10} = I'_{20}$ is counted off the accurate bearing α, which is determined by the formula

$$\operatorname{tg}\alpha = \frac{I'_2}{I'_1} = \operatorname{tg}\theta \text{ and } \alpha = \theta. \qquad (4.73)$$

Due to the fact that the length of the cable of the second framework differs from the length of the cable of the first framework $l_2 = l_1 + \Delta l$, current in the load of the cable of the second framework I'_{20} will change and it will stop $I'_{20} = I'_{10} + \Delta I$.

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Will appear the bearing error, which at small values $\Delta I/I'_{10}$ has quadratic character, and in accordance with formula (4.18) its maximum value is determined by the formula

$$\operatorname{tg}(\Delta a) \approx \frac{1}{2}\frac{\Delta l}{l'_{10}}. \qquad (4.74)$$

Differentiate of expression (4.71') in terms of ml_0 and after subdividing it into $4(I'_{10})^2$, let us find the maximum value of error $(\Delta a)_{\text{макс}}$:

$$\operatorname{tg}(\Delta a)_{\text{макс}} = \frac{1}{2}\,\frac{[(z_n^2-\rho_\Phi^2)(S^2-1)]\,\Delta ml_0 \sin 2ml_0}{(z_n+S\rho_\Phi)^2\cos^2 ml_0+(\rho_\Phi+Sz_n)^2\sin^2 ml_0}. \qquad (4.75)$$

Error is equal to zero independent of length l_0 with $z_n=\rho_\Phi$ or S = 1, which establish/installed earlier.

After equating zero derivative (4.75) in terms of ml_0, let us find the most unfavorable length of cable l_0, at which is obtained maximum possible error $(\Delta a)_{\text{макс макс}}$:

$$\cos 2ml_0 = \frac{(\rho_\Phi+Sz_n)^2-(z_n+S\rho_\Phi)^2}{(\rho_\Phi+Sz_n)^2+(z_n+S\rho_\Phi)^2}. \qquad (4.76)$$

After substituting (4.76) in (4.75), we will obtain

$$\operatorname{tg}(\Delta a)_{\text{макс макс}} = \left|\frac{(z_n^2-\rho_\Phi^2)(S^2-1)}{2(\rho_\Phi+Sz_n)(z_n+S\rho_\Phi)}\right|\Delta(ml_0). \qquad (4.77)$$

Using expression (4.72), let us determine the change in the phase, produced by a difference in the length of the pair of cables of one of the framework of Δml_0:

$$\Delta\varphi \approx \frac{(z_n + S\rho_\phi)(\rho_\phi + Sz_n)\,\Delta ml_0}{(z_n + S\rho_\phi)^2\cos^2 ml_0 + (\rho_\phi + Sz_n)^2\sin^2 ml_0}.$$

Maximum value $(\Delta\varphi)_{\text{макс}}$ is obtained when $ml_0 = (2n+1)\frac{\pi}{2}$ and it is equal

$$(\Delta\varphi)_{\text{макс}} = \frac{z_n + S\rho_\phi}{\rho_\phi + Sz_n}\Delta ml_0. \qquad (4.78)$$

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The ellipticity of magnetic field in goniometer (or the image of bearing on the cathode-ray tube of the two-channel radio direction finder), which is caused by a phase difference $(\Delta\varphi)_{\text{макс}}$, is determined from formula (4.23):

$$\left(\frac{A}{B}\right)_{\text{макс}} = 0{,}5\,(\Delta\varphi)_{\text{макс}} = 0{,}5\left(\frac{z_n + S\rho_\phi}{\rho_\phi + Sz_n}\right)\Delta ml_0. \qquad (4.79)$$

When $z_n = \rho_\phi$ or $z_n = \rho_\phi$ (i.e. S = 1), then $(\Delta a)_{\text{макс макс}} = 0$ and $(\Delta\varphi)_{\text{макс}} = \Delta ml_0$ independent of value Δml_0.

Most strongly manifests itself the dissimilarity of the lengths of cables when $z_n = 0$ or $z_n = \infty$. For these cases, taking into account that $S = K_0$, from the connection of the load of antenna, we will obtain

$$\left.\begin{aligned}
\operatorname{tg}(\Delta\alpha)_{\text{макс макс}} &= \left|\frac{1-K^2_{\text{бв}}}{2K_{\text{бв}}}\right|\Delta ml_0,\\
\left(\frac{A}{B}\right)_{\substack{\text{макс}\\ z_н=0}} &= 0{,}5K_{\text{бв}}\Delta ml_0, \quad \text{or}\\
\left(\frac{A}{B}\right)_{\substack{\text{макс}\\ z_н=\infty}} &= 0{,}5\frac{1}{K_{\text{бв}}}\Delta ml_0.
\end{aligned}\right\} \qquad (4.80)$$

If we consider that are obtained current resonance or voltages in antenna circuit, i.e., $z_н=\infty$, or $z_н=0$ (S = ∞, either S = 0), then we have

$$\operatorname{tg}(\Delta\alpha)_{\text{макс макс}} = \left|\frac{1-P^2}{2P}\right|\Delta ml_0$$

and

$$\left(\frac{A}{B}\right)_{\text{макс}} = \frac{1}{2}P\Delta ml_0,$$

← where $P=\frac{z_н}{\rho_ф}$, or $P=\frac{\rho_ф}{z_н}$ — a coefficient of travelling wave from the load of receiver.

Laws in this case the same, as (4.80), only $K_{\text{бв}}$ it is replaced by P.

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On Fig 4.28 are depicted designed by formulas (4.33) at values $\Delta ml_0=0{,}01$ and $\Delta ml_0=0{,}05$ the bearing errors Δα and the maximum ellipticity of magnetic field in goniometer A/B depending on KBV of the load of cable from receiver or antenna.

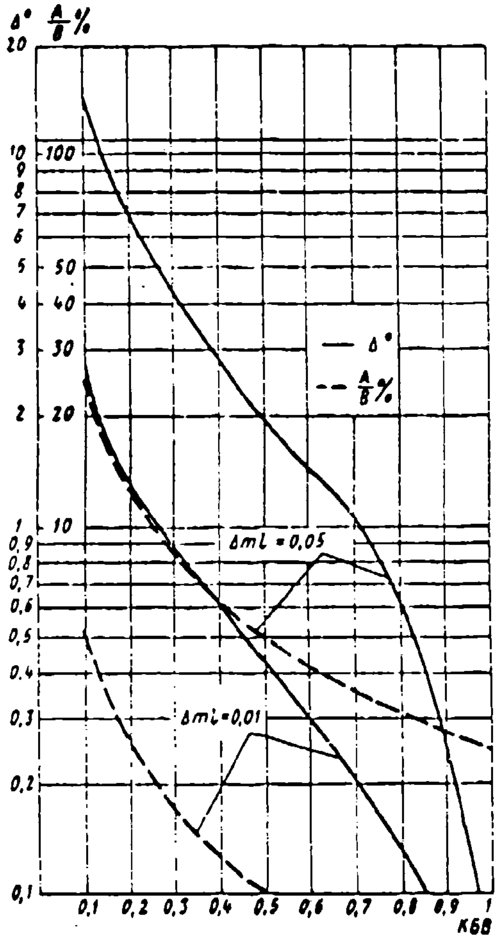

Fig. 4.28. Errors due to the dissimilarity of the electrical length of cables.

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From the curves Fig. 4.28 follows that, limiting error by value 1°, it is possible to allow KBV of load from 0.25 to 0.75 depending on the dissimilarity of the lengths of cable.

It is possible to show that the dissimilarity ρ_{ϕ} between the pairs of feeders is led to squared error whose maximum value $0{,}5\frac{\Delta\rho_{\phi}}{\rho_{\phi}}$, where $\frac{\Delta\rho_{\phi}}{\rho_{\phi}}$ — relative value of the dissimilarity of the wave impedance of cables. Usually $\frac{\Delta\rho_{\phi}}{\rho_{\phi}}$ they limit so that the maximum error would lie/rest within the required limits. Sometimes for the equalization ρ_{ϕ} of cables cable with high wave impedance they shunt by small amount of capacitance, selected experimentally.

On (4.78), it is possible to also calculate phase displacement of currents, which occurred as a result of the dissimilarity of the lengths of two cables in one pair, only Z_n and ρ_{ϕ} they are related in this case to single cable. Knowing phase displacement of currents in a pair, it is possible on (4.79) to determine the caused by this phase displacement antenna effect. Being given the permissible antenna effect due to the effect of the dissimilarity of the lengths of cable in a pair, is determined the permissible maximum dissimilarity of lengths.

We will examine system of two framework or of two pairs of antennas. If in antenna to system used more than four antennas, then instrument error does not exceed that which was designed by formula (4.77).

Effect of the dissimilarity of the electrical lengths of cables in by the circular antenna to system with acute/sharp directional characteristic.

We assume the sum-and-difference method of the reading of bearing.

2N - the number of antennas, which form direction-finding group,

E - voltage from any antenna on the output of the cell/element of addition.

Output potential of the cell/element of the addition of emf N of the antennas of subgroup with the identical lengths of the cable

$$E_1 = EN.$$

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Output potential of the cell/element of the addition of emf N of the antennas of the subgroup where one antenna has a cable with a differing from others length of:

$$\dot{E}_s = E\left[(N-1) + e^{j\Delta m l}\right]$$

or

$$\dot{E}_s = E\left[(N-1)^2 + 1 + 2(N-1)\cos(\Delta m l)\right]^{1/2} \times$$
$$\times e^{j \operatorname{arc\,tg} \frac{\sin(\Delta m l)}{N-1+\cos(\Delta m l)}}.$$

With the reading of bearing, appear the bearing error Δ and the ellipticity of image A/B.

The corresponding to error angle on cathode-ray tube will be

$$\Delta\alpha = 0{,}5 \operatorname{arc\,tg} \frac{\sin(\Delta m l)}{N-1+\cos(\Delta m l)}.$$

Bearing error with consideration scaling factor $k_{\text{п}} = \frac{2\cdot b_{\text{э}}}{\lambda}$ $(2b_{\text{э}} -$ the separation of the equivalent pair of antennas, which replaces direction-finding group) is determined by the formula

$$\Delta = \frac{\Delta\alpha}{k_{\text{п}}}.$$

It is possible to show that the ellipticity of image is designed from formula (see § 8.7)

$$\frac{A}{B} = \frac{\left|\sqrt{(N-1)^2 + 1 + 2(N-1)\cos(\Delta m l)} - N\right|}{\left|\sqrt{(N-1)^2 + 1 + 2(N-1)\cos(\Delta m l)} + N\right|}.$$

Limiting error by value $\Delta \leqslant 0.1^\circ$, we obtain for the permissible value $\Delta ml < 2\cdot 0{,}1 \frac{1}{57} \frac{2\pi b_0}{\lambda} N$ or $\Delta ml < 0{,}02 \left(\frac{b_0}{\lambda} N\right)$.

Thus, when $\frac{b_0}{\lambda}=1$, N = 4 requirement $\Delta ml < 0{,}08$, or $\Delta l \leqslant 0{,}01\lambda$ should be satisfied.

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Analogously are determined tolerances to manufacture of the lines of time delays in the antenna commutator.

Effect of the dissimilarity of the electrical lengths of cables in a radio direction finder with the cyclic measurement of phase in high frequency.

We assume use by the circular antenna of system with switching of antennas for determining bearing.

The supplementary phase of emf in one of the antennas $\varphi_0=\Delta ml$ will lead to a sinusoidal increment in the phase with amplitude [1.17]

$$\zeta \doteq \frac{2}{\pi} \sin\left\{\frac{\pi}{N} \Delta ml\right\},$$

where N is a number of antennas.

Greatest affect an increment in the phase ζ exerts itself when the sinusoid of an increment in the phase is shifted on angle on π/2 relative to the fundamental sinusoid of a change in the phase with the frequency of commutation. in this case, the bearing error is designed from the formula

$$\mathrm{tg}\,\Delta_{\text{макс}} = \frac{\zeta}{B} \quad \text{or} \quad \Delta_{\text{макс}} \approx \frac{\zeta}{B},$$

where B = 2πb/λ; b - a radius by the antenna of system.

Let $B_{\text{мин}} = 1$ and N = 12. Then for limitation $\Delta_{\text{макс}} < 0{,}1°$ must be

$$(\Delta ml) < \frac{0{,}1}{57}6, \quad \text{or} \quad (\Delta ml) < 0{,}0106.$$

For determining the total instrument error of radio direction finder, it is necessary to accumulate component instrument errors. Since the component errors have different dependences on azimuth (but sometimes and on frequency), most it is correct to rate/estimate the average quadratic values of component errors and then, assuming that the errors are independent, to calculate total mean square error as square root of the sum of the dispersions of component errors.

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It is natural that of radio direction finder with rotary antenna by system the complete instrument error is less than in radio direction finder with motionless antenna by system. The instrument arror of stationary radio direction finder is less than movable or ship (aircraft).

Instrument arror can be to a certain degree taken into account by the calibration of radio direction finder (see Chapter 10).

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Chapter 5.

EFFECT OF LOCALITY AND ENVIRONMENT.

5.1. Character of the effect of locality and environment.

As it was shown earlier, with the aid of radio direction finder is determined the orientation of the equiphase surfaces of electromagnetic field, emitted by the oriented radio transmitter. At a great distance from transmitter on the limited section of arrangement/permutation by the antenna of the system of the radio direction finder of the projection of equiphase surfaces on ground, that are the concentric circumferences of a large radius with center at the point of the location of transmitter, are converted into straight lines, perpendicular to direction in radio station. Direction in the transmitting radio station, i.e., line of bearing, is determined from the perpendicular from center by the antenna of system to equiphase surface.

Due to influence of environment with the antenna of the system of the radio direction finder of the projection of equiphase surfaces, they are distorted, and then appears error with direction finding. The distortion of equiphase surfaces and error they can be caused by the heterogeneity of soil and area relief near by the antenna of the system of radio direction finder (transition from humid soil to dry, from sea to dry land and so forth). Furthermore, the different metallic and current-conducting installations and object/subjects (antennas, the locked ducts, hangars, trees, etc.) affect the work of radio direction finder, since they create the fields of reradiation, which distort the orientation of equiphase surfaces and the calling errors and the diffuseness of bearing. So that the locality and the local installations would not affect the antenna system of ground-based radio direction finder, it place, so that to the system would not exert effect the heterogeneity of soil and relief, and also the current-conducting object/subjects.

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In spite of this, the locating at a great distance from radio direction finder (to several kilometers) current-conducting objects (structures, antenna installations, forest and so forth) are created in the sum of error with direction finding. Since the parameters of soil, and also the character of the emission/radiation of the distant environment can change depending on weather, the polarizations of the incident wave and other conditions, effect of distant environment does not remain time-constant. Errors due to the distant environment are of a random character and cannot be taken into account by calibrating the radio direction finder. They reduce the operating accuracy of radio direction finder. The greater the separation of antennas and is acute/sharper the antenna radiation pattern of the system of radio direction finder, the lesser the remote return emitters it operates on it and the lesser the random errors due to the effect of the distant environment.

In certain cases, for example during the installation of radio direction finder on ship or aircraft, it is not possible to avoid the effect of metallic object/subjects; then they occur of the errors, considered with direction finding in the form of the corrections, called radio beam deviation.

5.2. The shore effect

If near radio direction finder passes the shore line, which separate/liberates sea from dry land (or dry lands from sea), then with direction finding are possible the errors, called errors due to coastal effect.

The theory of radiowave propagation under these conditions is developed by V. A. Fok, M. A. Leontovich, G. A. Greenberg and Ye. L. Feynberg [5.2].

Usually transition from sea to dry land or vice versa is accompanied by the presence, in the first place, of slope of shore and, secondly, by a change in the electrical parameters of medium during the intersection of shore line. Both these reasons create errors with direction finding.

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Total error due to coastal refraction or the coastal effect

$$\alpha = \alpha_p + \alpha_n,$$

where α_p is due to the inequality of surface (slope), α_n - dissimilarity of the electrical parameters of soil on both sides from shore line.

In the mathematical analysis of the theory of coastal refraction, are utilized complex bulky conclusions. Let us give only obtained results [1.3, 5.2, 5.5].

Effect of the heterogeneity of soil.

Let at point O (Fig. 5.1) above sea is be located the emitter, and at point C above land - radio direction finder. Let us connect points O and C and direct axis OX of coordinate system along line OC. Let us designate: g - distance from emitter to radio direction finder; X_n is a path length above the real; ξ is a distance from C to shore line on perpendicular to shore line; 0 - the angle of incidence in the wave on shore line; ε and σ are the parameters of dry land (ε - the module/modulus of composite dielectric constant, σ - conductivity).

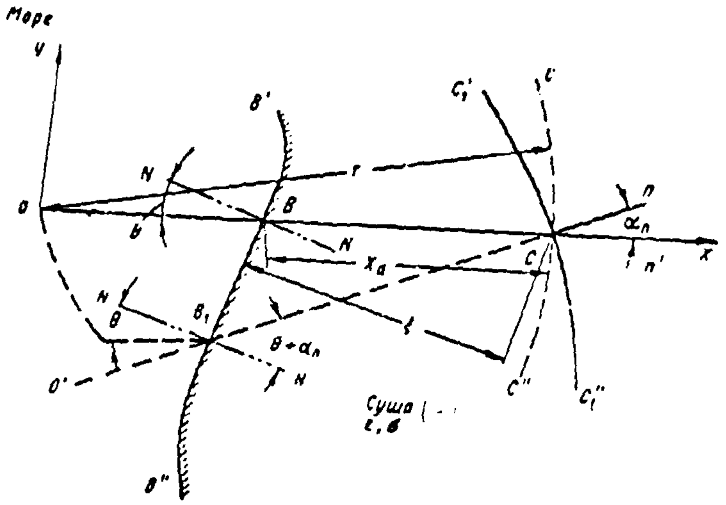

Fig. 5.1. Refraction of radio waves by shore line.

Key: (1). Sea. (2). Dry land.

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Electromagnetic field along the surface of propagation (dry land or seas) is characterized by the function of weakening W, which is the complex quantity, which depends on the parameters of the medium, above which is propagated the wave. The module/modulus of this value characterizes the decrease of the amplitude of the strength of field with distance, the argument of the function of weakening determines the supplementary phase of wave relative to the phase in free space ($2\pi/\lambda\ r$).

If an increment in the phase after the intersection with the wave of the line of shore was constant, not depending on direction of propagation, then the lines of equal phases would be circumference with center at point 0 of the location of the emitter (dotted line C'C" on Fig. 5.1). The radio bearing, which is determined from perpendicular On' to the line of equal phases, would coincide with direction in radio transmitter.

It is known that the wave front, which is propagated above the ideally conducting surface (by sea), is perpendicular to this surface. Above the semi-conducting earth/ground the wave front of interface it is sloped forward the greater, the lesser the ground conductivity. At a height of several wavelengths the distortion of wave front disappear and it coincides with the sphere, which has center at the point of the location of emitter. During the transition of the wave through the shore line, which divides mediums with different conductivity (sea and dry land), the wave front of surface must undergo change from the normal to inclined. Simultaneously with this occurs an increment in the supplementary phase of the function of weakening.

During radiowave propagation perpendicular to shore line slope deviation in the wave front and an increment in the supplementary phase occurs simultaneously on all sectors of the front of wave; the direction of the line of the equal phases C'C" will not change. If wave intersects shore line at certain angle of $0 \neq 0°$, then the individual sections of wave front pass from sea to dry land not simultaneously and an increment in the phase in different sectors of the front of wave occurs also not simultaneously, but with the intersection by it of shore line.

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As a result of this the line of the identical phases is distorted and assumes the form of solid line $C'_1C''_1$ (Fig. 5.1). Perpendicular Cn to the line of identical phases does not coincide with direction in transmitter. Appears bearing error [illegible] depending on the direction of the motion of wave of relatively shore line B'B". Wave seemingly undergone refraction - its path seemingly stops OB_1C. During the further advance of wave with the establishment of the necessary slope/inclination of the electric field above the earth/ground, occurs phase compensation in all sectors of the front and gradually is restored the direction of the line of equal phases, perpendicular to the straight line OC. Thus, on certain distance from shore line the error of coastal refraction disappers.

Study of coastal refraction showed that the errors of coastal refraction is detected while direction finder is located from shore line at a distance, much smaller than $0{,}318\lambda|\varepsilon|$.

Furthermore, as this follows from that which was presented, the error in question must not be observed with climb of several wavelengths.

In [5.2] obtained common expression for an error from coastal refraction under the following assumptions: plane wave intersects the rectilinear coastal feature of infinite length; on both sides from this line of medium, have the different parameters, the transition from one parameter to another occurring smoothly; the width of transient zone is small in comparison with wavelength.

Is derived a series of formulas for special cases.

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For the case when radio direction finder is located on dry land far beyond the limits of transient zone and transmission occurs from sea, is obtained formula

$$\alpha_n = -\frac{\operatorname{tg}\theta}{\sqrt{2\pi m_c \Delta_n}} \sqrt{\eta_c} \sin\left(\frac{\pi}{4}+\frac{\beta_p}{2}\right)\sqrt{1-\frac{\Delta_n}{r}}. \quad (5.1)$$

where

$$\eta_c = \frac{1}{\sqrt{\varepsilon^2+\left(\frac{2\sigma\lambda}{c}\right)^2}};$$

$$\beta_p = \operatorname{arc\,tg}\left(\frac{2\sigma\lambda}{\varepsilon c}\right),$$

c - the speed of light.

Minus sign before the formula designates, that the direction of propagation after the intersection of shore line approaches a standard. If we interchange the position radio transmitter and direction finder, then error will become positive.

Figures 5.2 gives nomogram for calculation α_n for distances from shore line according to perpendicular to it, equal to $\xi = \lambda/3$, $\xi =$

$= \lambda$ and $\varepsilon - 5\lambda$ for two waves 300 and 600 m and for three varieties of soils. The parameters of soil in this figure are designated in unity CGSE. Being given angle of incidence θ and ratio r/ε, on the left-handed curves finds the point, from which it is necessary to conduct the horizontal line to intersection from one of the vertical lines on which is counted off the error.

The comparison of those who were obtained up to the development of the theory of the experimental observed data of the errors of coastal effect and theory as a result of the absence of the comprehensive data on measuring conditions is difficult. One should, however, note that the sign and the order of magnitude of the observed errors will agree well with theory.

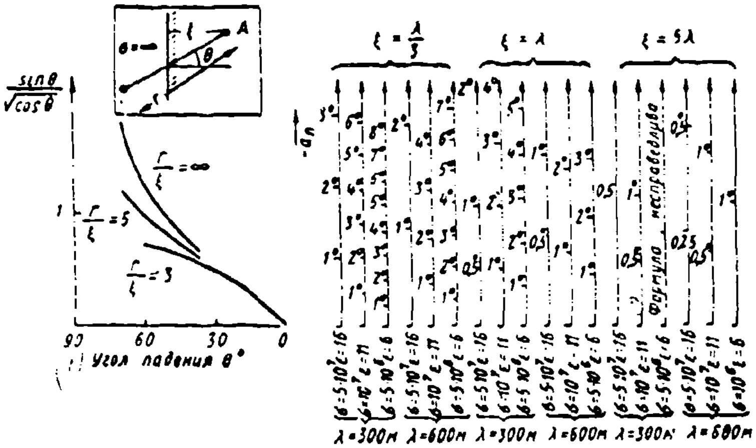

Fig. 5.2. Nomogram for calculating error due to coastal effect.

Key: (1). Angle of incidence θ°.

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If during the propagation of wave the section of dry land is shorter than marine, then angular error is less with direction finding from the ship, than with direction finding from shore. Depending on angle of incidence, the error $\alpha_{н}$ has approximately semi-circular nature with zero along perpendicular to shore line.

The errors, analogous to coastal refraction, can be, also, when on path of motion wave does not intersect shore line, but there are heterogeneities of the electrical parameters or relief of soil. Investigations show that the small heterogeneities of soil do not manifest themselves by noticeable shape, if observations are conducted outside the range of the distortion of the phase response of field, and they can lead to the considerable disturbances of the character of field and the bearing errors in the ranges, included by these distortions.

The electrical parameters of soil change with a change in the atmospheric conditions. Therefore the errors, caused by the heterogeneity of the parameters of soil, do not remain time-constant and cannot be considered them during the operation of radio direction finder by the pre-check of radio direction finder. The value of these errors must be restricted by appropriate selection of place for the installation of radio direction finder. Before its installation it is necessary to investigate the parameters of soil of area by a radius, by the approximately equal to 1/3-1 maximum wavelengths of direction finding, around site of installation by the antenna of the system of radio direction finder. Site of installation can be considered satisfactory, if the spread of ground conductivity within the limits of the mentioned area is small.

Is experimentally establish/installed the linear dependence between the fluctuations of earth conductivity and the mean error of spaced-antenna direction finder. During the fluctuations of specific conductivity 1-4 in a radius 100-120 m, the maximum bearing error on short waves has value ±1°.

Separately must be checked ground conductivity directly in the sites of installation of masts, since the dissimilarity of soil under masts causes the dissimilarity of the electrical parameters of the antennas and instrument errors at direction finding.

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The metallization of soil in site of installation by the antenna of system of radio direction finder, Reamer by means of the laying of wire gauze (see § 6.3), it leads to the fact that the dissimilarity of ground conductivity under grid manifests itself to much smaller degree.

Effect of the inequality of soil.

It is assumed, that along the rectilinear shore of infinite length is a slope with constant rate of rise.

Let us designate: γ_0 - the rate of rise of slope in its central part; l extent of slope; ζ_0 - the maximum value of lift; ξ is a distance on perpendicular to shore line from observation point to shore line, from observation point to the shore line where the slope begins; ξ has minus sign, if observation point is located in front of slope (at sea), and plus sign, if observation point is located behind slope (on the shore).

On Fig. 5.3, is represented the section by the vertical plane of the medium above which are propagated the radio waves, intersecting shore line.

Figures 5.4 gives the curve/graphs of the maximum errors of direction finding α_p (into rad), which are obtained, when wave intersects the slope, which is pulled along the rectilinear coastal feature of infinite length. Errors are given depending on θ for the different $m\xi$, for the case $\gamma_0 < \frac{1}{2\pi}\sqrt{\frac{\lambda}{T}}$, $\left(m=\frac{2\pi}{\lambda}\right)$.

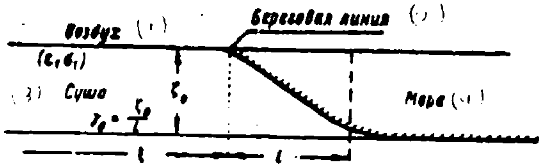

Fig. 5.3. Airfoil/profile of medium during the intersection of shore line.

Key: (1). Air. (2). Shore line. (3). Dry land. (4). Sea.

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From the curves Fig. 5.4 it is evident that at equidistance from slope the error when direction finder is located in front of slope at sea ($m\xi$ negative), is approximately two times more than when direction finder is located behind slope on dry land ($m\xi$ positive).

For example, with distance $\xi = \lambda/2\pi$, $\theta = 45°$ and $\gamma_0 = 0.2 = 12°$ error of $a_p = 2$-$3°$ in front of slope and $a_p = 1$-$1.5°$ under the same conditions is behind of slope. Error increases with an increase of the wavelength and depending on θ has quadratic character for $m\xi >$

-1.5; for $m\xi < -1.5$ dependence of error on θ becomes more complex (Fig. 5.4).

It is experimentally establish/installed that the errors of coastal effect most powerfully are developed of the waves with a length of 500-1000 m. On the waves, greater than 3000 m, they are less than 1°. On wave 500 m in several meters of shore, is observed the error at $\theta = 70°$ into 3-4°. During the elongation of wave from 500 to 2600 m the error decreases from 3.2° to 1.4°. On the short waves of error from coastal effect, they decrease, but they are not systematized, on VHF there are not.

Calculations and practical data show that the radio bearing, which undergoes coastal effect, is sufficiently flowed, if electromagnetic wave intersects shore at an angle, greater than 20° ($\theta \leq 70°$), and if distance X_{n} on the path of propagation from shore line to direction finder is more than one or several waves.

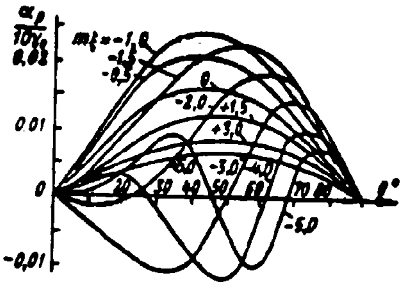

Fig. 5.4. Errors of relief.

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For coastal radio beacons and radio direction finders on charts, they usually note the zone of reliable direction finding, within limits of which it is possible to rely on a small manifestation of coastal effects.

5.3. Effect of the adjacent object/subjects on radio direction finder.

Effect on the direction finding of the metallic object/subjects of those located near by the antenna of the system of radio direction finder, manifests itself in the fact that can appear the bearing errors and sometimes diffuseness in its reading (blurring the minimum of audibility in auditory radio direction finder, appearance on the cathode-ray tube of the two-channel automatic direction finder of ellipse instead of the line and so forth).

Is explained this as follows. The incoming electromagnetic wave induces emf in the metallic object/subjects, located near by the antenna of the system of radio direction finder. These emf's create in metallic object/subjects, the currents which form their electromagnetic fields, called the fields of reradiation. The latter operate on the antenna system of radio direction finder together with ground field of transmitter.

The emf induced with the field of return emitter, it is possible to decompose on two components - that cophasal with emf, induced with ground field, and differing from it in phase in $\pi/2$. The first of them creates error with direction finding, the second causes in essence the diffuseness of reading. Emf from the field of reradiation, which is characterized by in phase in $\pi/2$ from emf of ground field, it is possible to compensate for the same means which are applied for the compensation for emf from antenna effect, which differs in phase in $\pi/2$ from emf of the directed system (§ 4.3). If this compensation is realized, then remains the action only of one field component of reradiation, cophasal with ground field and which creates only error with direction finding.

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If in frame radio direction finder for determining one-sided bearing (side of radio station) is utilized the superposition method of the diagram of the reception/procedure of the omnidirectional antenna on the diagram of the reception/procedure of the framework, then the field of reradiation can lead to the incorrect reading of side. The reason for this consists in the fact that in the near zone of reradiation is disrupted phase relationship of magnetic component of electromagnetic field, which operates on the framework, and by the electrical component, which operates on antenna.

Let us examine in more detail the action of the field of reradiation on radio direction finder with cosinusoidal directional characteristic. For simplicity we assume that as by the antenna of system is used the rotatable loop or the pair of the spaced antennas

(Fig. 5.5).

To the rotatable loop of radio direction finder R with effective height h_p at an angle p the initial reference line 00, approaches the electromagnetic wave of transmitter (ground field) with intensity E and with normal polarization (vector E in the vertical plane of propagation).

The phase of field let us count off relative to point R. At point A at an angle ψ to initial reference line at a distance d from the center of the framework, is located return emitter with effective height $h_{\text{он}}$ by resistor/resistance $Z_{\text{он}}=z_{\text{он}}e^{-j\varphi_0}$ and by directional characteristic F (θ_0, θ), where θ_0 is an angle of the direction of the maximum of radiation pattern of return emitter with initial reference line, that characterizes the orientation of return emitter, θ - the angle of any direction in question with the direction of initial reference line. We assume that the angle of the slope of a front of wave $\beta = 0$.

In return emitter it is induced by emf

$$\dot{E}_{\text{он}}=E\,h_{\text{он}}F(\theta_0,\,\psi_0)\,e^{j\varphi_0}.$$

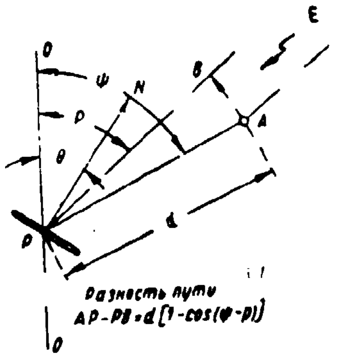

Fig 5.5. Action of return emitter.

Key: (1). difference in the path.

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In expression for directional characteristic instead of the angle θ, is introduced the angle $\psi_0 = \dot{p} - \psi$ between directions in the oriented radio station and in return emitter from the location of the framework of radio direction finder; a phase difference ϕ_2 is caused by the character of emitter (locked, extended), by the location of return emitter relatively by the antenna of the system of radio direction finder and by direction of propagation.

When ground field has direction of propagation $p = \psi + \pi/2$, then a phase difference $\phi_2 = \phi_{20}$ depends only on the character of return emitter.

For any direction $p \neq \psi + \pi/2$ (Fig. 5.5)

$$\varphi_2 = \frac{2\pi}{\lambda} d \cos \psi_0 + \varphi_{20},$$

where d = AP is a distance between the return emitter and the antenna the system of radio direction finder.

Current in return emitter [1].

$$I_{0n} = \frac{\dot{E}_{0n}}{Z_{0n}} = \frac{E h_{0n} F(\theta_0, \psi_0)}{Z_{0n}} e^{j(\varphi_1+\varphi_2)} = I_{0nm} e^{j(\varphi_1+\varphi_2)}.$$

FOOTNOTE [1]. In the case of nonuniform distribution of current along return emitter h_{0n}, Z_{0n}, E_{0n} and I_{0n} they are related to maximum current in return emitter. ENDFOOTNOTE.

Return emitter creates in the location of the framework of radio direction finder its field of reradiation, we assume that with the normal polarization

$$\dot{E}_{\text{ои}} = aI_{\text{ои}} F(\theta_0, \psi) e^{j\varphi_1} = aI_{\text{ои}\, m} F(\theta_0, \psi) e^{j(\varphi_1+\varphi_2+\varphi_3)};$$

the supplementary phase Φ_3 depends on the alignment of the return emitter and framework of radio direction finder and on the character of return emitter. If $d << \lambda$, then $\Phi_3 = \Phi_{30}$ is determined by the character of return emitter.

For any direction whether large d (Fig. 5.5)

$$\varphi_3 = \pm \frac{2\pi}{\lambda} d + \varphi_{30}.$$

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We can write for an electrical component electromagnetic field of the return emitter

$$\dot{E}_{\text{ои}} = E_{\text{ои}\, m} e^{j\varphi} = E_{\text{ои}\, m} \cos\varphi + jE_{\text{ои}\, m} \sin\varphi = E'_{\text{ои}} + jE''_{\text{ои}}, \tag{5.2}$$

where $E_{\text{ои}\, m} = kE;$

$$k = \frac{a h_{\text{он}} F(\theta_0, \psi_0) F(\theta_0, \psi)}{z_{\text{он}}}; \qquad (5.3)$$

$$\varphi = \varphi_1 + \varphi_2 + \varphi_3. \qquad (5.4)$$

Field component of return emitter, cophasal with the field of transmitter, will be

$$E'_{\text{он}} = k E \cos \varphi. \qquad (5.5)$$

Field component of return emitter, which differs in phase on $\pi/2$ from the field of transmitter, is expressed

$$E''_{\text{он}} = k E \sin \varphi. \qquad (5.6)$$

The first component is led to bearing error, the second component causes mainly diffuseness in the reading of bearing. With $\phi = 0$ or $\phi = \pi$ occurs the maximum of error.

For research on the effect of return emitter, it is necessary to determine its field $E_{\text{он}}$, i.e. to find values of k and ϕ [see (5.3) and (5.4)].

Coefficient k depends on form and the alignment of return emitter, and also on the ratio of its natural frequency to the frequency of transmitter. Phase ϕ depends on the character of emitter and its resistor/resistance, determined by the ratio of the natural frequency of emitter to the frequency of direction finding (component phases $\phi_1 + \phi_{20} + \phi_{30}$), and also on the mutual location of return emitter and by the antenna of the system of radio direction finder $(\phi_2 + \phi_3 - \phi_{20} - \phi_{30})$. This phase difference is equal to $2\pi d/\lambda\ (1 \pm \cos\psi_0)$.

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If around by the antenna of the system of radio direction finder are several return emitters, then for determining their action it is necessary for the location of the directional antenna of radio direction finder to accumulate the fields of all return emitters taking into account their mutual phases.

Since the phases of the fields of emitters in the general case depend on the direction of the arrival of ground field, to store/add up the fields of return emitters is necessary for each separate direction. According to the characteristics of the total field of all return emitters for each direction, are designed the error and the diffuseness of bearing.

Let us find expression for a bearing error in the general case. Let the standard to the plane of the framework form angle θ with initial reference line. If in radio direction finder is not provided the compensation for antenna effects, then within the framework of radio direction finder it is induced by emf

$$\dot{E}_p = Eh_p[\sin(p-\theta) + k\cos\varphi\sin(\psi-\theta) + j\,k\sin\varphi\sin(\psi-\theta)]. \qquad (5.7)$$

The amplitude of stress E_p is determined by the formula

$$E_p = \sqrt{[\sin(p-\theta)+k\cos\varphi\sin(\psi-\theta)]^2+[k\sin\varphi\sin(\psi-\theta)]^2}\;Eh_p. \qquad (5.8)$$

Emf within the framework is equal to zero not at which values of angle θ. Bearing is counted off on the minimum of audibility. For determining the value of bearing θ = q with which occurs the minimum of the audibility of radio station, one should equate zero derivative $\frac{dE_p}{d\theta}$.

Instead of E_p we examine the expression, proportional E_p:

$$G = \sqrt{[\sin(p-\theta)+k\cos\varphi\sin(\psi-\theta)]^2+[k\sin\varphi\sin(\psi-\theta)]^2} \qquad (5.9)$$

or

$$G^2 = \sin^2(p-\theta) + k^2\sin^2(\psi-\theta) + 2k\cos\varphi\sin(p-\theta)\sin(\psi-\theta). \qquad (5.10)$$

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Let us equate zero derivative of G^2 in terms of θ and is simultaneously considered, what on the minimum θ = q and p - θ = p -

q = f is a correction to radio bearing [1],

$$\frac{d(G^2)}{d\theta}=2\sin f\cos f+2k^2\sin(\psi-p+f)\cos(\psi-p+f)+$$
$$+2k\cos\varphi\sin f\cos(\psi-p+f)+$$
$$+2k\cos\varphi\cos f\sin(\psi-p+f)=0,$$

whence correction to bearing f is determined by the expression

$$\operatorname{tg}2f=\frac{2k\cos\varphi\sin\psi_0+k^2\sin2\psi_0}{1+2k\cos\varphi\cos\psi_0+k^2\cos2\psi_0}. \qquad (5.11)$$

FOOTNOTE [1]. With examination of the effect of return emitters, we speak about correction to bearing, as is customary in the theory of deviation. Bearing error has a sign, reverse/inverse to the sign of correction. Subsequently for the calculation of errors, we use the formulas, obtained for corrections. ENDFOOTNOTE.

Angle q_1 at which is counted off the bearing, is determined by the formula

$$\operatorname{tg}2q=\frac{\sin2p+2k\cos\varphi\sin(p+\psi)+k^2\sin2\psi}{\cos2p+2k\cos\varphi\cos(p+\psi)+k^2\cos2\psi}.$$

At the low value of k, the correction will be

$$f\approx\operatorname{tg}f=k\sin\psi_0\cos\varphi. \qquad (5.12)$$

When k > > 1, is oriented actually return emitter. The calculated bearing independent of a change in the angle of radio station p remains constant and equal to ψ. Bearing deviation

$$f=p-\psi=\psi_0.$$

When k = 1, deviation from expression (5.11) is obtained equal by $f = \Psi_0/2$ or $f = \pi/2 + \Psi_0/2$.

Formulas (5.11) and (5.12) are used also for goniometric and two-channel systems.

From expression (5.12) it follows that if the phase of field of return emitter Φ artificially is changed from 0 to 2π, then bearing error varies from maximum positive to maximum negative value.

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Therefore the designed during this artificial change in the phase difference Φ average/mean bearing has an error the lesser, than more frequent they are taken readings and than less k.

The imposition of the images of bearings on the cathode-ray tube of two-channel automatic or single-channel with the revolving goniometer radio direction finder during special change from 0 to 2π of the relative phases of fields of fundamental and reradiations makes it possible to recognize at multiple-pronged field bearings for separate emission/radiations [5.15, 10.4]. Is explained this by following. In two-channel radio direction finder with direction finding two- and of triradial field, the ellipses of the images of separate bearings during an artificial change in phase relationship of fields form parallelogram or the parallelepiped whose sides correspond to bearings to separate field component. In single-channel radio direction finder with the revolving goniometer, the total output potentials of receiver during a change in the relative phase difference of fields from 0 to 2π have the nonmoved minimums, which determine bearings to field component.

Utilizing this property, in VHF range it is possible to achieve a decrease in the errors from return emitters, revolving antenna system with cosinusoidal directional characteristic in circumference with the radius, equal to wavelength or large it, and counting off averaged bearing [5.13]. In the range of short waves, it is suggested to obtain the images of bearings simultaneously in the fields of fundamental and return emitters by application/use several spread in the distance, commensurable with wavelength, automatic two-channel radio direction finders. These radio direction finders have the common/general/total indicator cathode-ray tube on which is obtained the parallelogram, which characterizes the directions of both fields [10.4].

When a phase difference $\varphi = 0°$ or $\varphi = \pi$, from (5.11) follows that

$$\mathrm{tg}\,2f = \frac{\pm 2k \sin\psi_0 + k^2 \sin 2\psi_0}{1 + k^2 \cos 2\psi_0 \pm 2k \cos\psi_0}$$

or

$$\mathrm{tg}\,f = \frac{+k \sin\psi_0}{1 \pm k \cos\psi_0}. \qquad (5.13)$$

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Sign (+) corresponds to a phase difference 0°, sign (-) corresponds to a phase difference π.

If we into formula (5.12) substitute $\psi_0 = q + f - \psi$, then we will obtain

$$\sin f = k \sin(q - \psi) \cos\varphi, \qquad (5.14)$$

where q is a radio-course angle (counted off bearing).

Formula (5.14) can be obtained directly from condition for the reading of the bearing:

$$[E \sin(p - \theta) + E'_{ов} \sin(\psi - \theta)]\, h_p = 0. \qquad (5.15)$$

For $\varphi = 0$ and $\varphi = \pi$ maximum value of error occurs when $\cos(p - \psi) = k$, i.e., $p - \psi = \psi_0 \approx \pi/2$,

$$\mathrm{tg}\,f_{макс} = \frac{\pm k}{\sqrt{1 - k^2}}.$$

If $\varphi = \pi/2$, from bearing error on the basis of formula (5.11) will be

$$\mathrm{tg}\,2f = \frac{k^2 \sin 2\psi_0}{1 + k^2 \cos 2\psi_0}. \qquad (5.16)$$

In this case the maximum of the error is obtained with $\cos 2\psi_0 = k^2$, or $\psi_0 \sim \pi/4$

$$|f_{\text{макс}}| \approx \frac{1}{2}\frac{k^2}{\sqrt{1-k^4}}. \qquad (5.17)$$

Maximum deviation for $k < 1$ has much smaller value when $\phi = \pi/2$, than with $\phi = 0$ or $\phi = \pi$.

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The examined cases of constant phase displacement $\phi = 0$, $\phi = \pi/2$ and $\phi = \pi$ independent of direction p are possible, if $d/\lambda\phi\phi < < 1$, i.e., mainly on average/mean and long waves. Phase displacement $\phi = 0$ corresponds also to the presence in the radio direction finder of elements of the compensation for antenna effects.

In the general case the expression for the diffuseness of the minimum of A/B (relation of the semi-axes of the ellipse of the field of goniometer or the ellipse of the image of bearing on the cathode-ray tube of two-channel radio direction finder) is bulky. When $\psi_0 = 90°$, i.e., direction in return emitter composes 90° with direction in radio station, then of (5.7) we will obtain

$$\frac{A}{B} \approx \sqrt{\frac{1+k^2}{2} - \frac{1}{2}\sqrt{1+2k^2\cos 2\varphi + k^4}}. \qquad (5.18)$$

In the case of $k < 1$

$$\frac{A}{B} = \frac{k \sin\varphi \sin\psi_0}{1+2k\cos\varphi\cos\psi_0};$$

when $\phi = \pi/2$

$$\frac{A}{B} = k\sin\psi_0.$$

With transition from cosinusoidal antenna radiation pattern of the system of the radio direction finder to of acute/sharper error due to the effect of the fields of reradiation they decrease. This decrease the greater, the acute/sharper the radiation pattern, i.e., is more the separation of antennas.

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if the separation between antennas 2b such, that it cannot be counted $2\pi b/\lambda << 1$ and direction finding is conducted on the minimum, then bearing error at any value ϕ is determined from the equality

$$\sin\left(\frac{2\pi b}{\lambda}\sin f\right) + k\cos\varphi \sin\left[\frac{2\pi b}{\lambda}\sin(\psi_0 + f)\right] = 0$$

or approximately, at the low values of f,

$$\sin f \approx f = \frac{k\cos\varphi\sin\left(\frac{2\pi b}{\lambda}\sin\psi_0\right)}{\frac{2\pi b}{\lambda}\left[1 + k\cos\varphi\cos\psi_0\cos\left(\frac{2\pi b}{\lambda}\sin\psi_0\right)\right]}. \quad (5.19)$$

With $\varphi = 0$ or $\varphi = \pi$

$$\sin f \approx f = \frac{\pm k\sin\left(\frac{2\pi b}{\lambda}\sin\psi_0\right)}{\frac{2\pi b}{\lambda}\left[1 \pm k\cos\psi_0\cos\left(\frac{2\pi b}{\lambda}\sin\psi_0\right)\right]}. \quad (5.19')$$

Sign (+) in (5.19') corresponds to a phase difference between E and E_{0H} $\varphi = 0$, sign (-) corresponds $\varphi = \pi$. For the low values of k, formula (5.19) is simplified:

$$f = \frac{k\cos\varphi}{\frac{2\pi b}{\lambda}}\sin\left[\frac{2\pi b}{\lambda}\sin(p - \psi)\right]. \quad (5.20)$$

Formula (5.19) is obtained on the assumption that the action of the nonphase field of return emitter is compensated for. If the compensation for nonphase field is absent and the reading of bearing is realized on the minimum of audibility or along the transverse of the image of bearing on the screen of the cathode-ray tube of two-channel radio direction finder, then error is determined by the expression

$$\sin / \approx \frac{k \sin\left(\frac{2\pi b}{\lambda} \sin \psi_0\right)\left[\cos \varphi + \right.}{\frac{2\pi b}{\lambda}\left[1 + k^2 \cos^2\left(\frac{2\pi b}{\lambda} \sin \psi_0\right)\cos^2\psi_0 + \right.} \rightarrow$$

$$\rightarrow \frac{\left. + k \cos\left(\frac{2 \cdot b}{\lambda} \sin \psi_0\right) \cos \psi_0\right]}{\left. + 2k \cos \psi_0 \cos\left(\frac{2\pi b}{\lambda} \sin \psi_0\right) \cos \varphi\right]}.$$

For $\phi = 0$ and $\phi = \pi$ is obtained the formula (5.19').

For $\phi = \pi/2$

$$\sin / \approx \frac{1}{2} k^2 \frac{\lambda}{2\pi b} \sin\left(2 \frac{2\pi b}{\lambda} \sin \psi_0\right) \cos \psi_0.$$

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The value of error considerably decreases and its maximum is observed when $\psi_0 \approx \frac{\lambda}{8b}$, rad. With a small separation of antennas, the expression for an error coincides with (5.16).

On Fig. 5.6, are depicted the errors due to the effect of return emitter of system of two omnidirectional spaced antennas depending on the angle between directions in return emitter and ground field $\psi_0 = p - \psi$ with k = 0.5 and different 2b/λ.

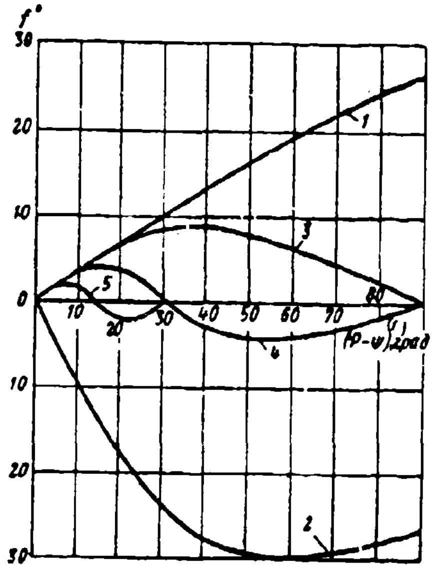

Fig. 5.6. Errors due to the effect of return emitter (k = 0.5): 1 - cosinusoidal characteristic ϕ = 0°; 2 - cosinusoidal characteristic ϕ = 180°; 3 - 2b/λ = 1, ϕ = 0; 4 - 2b/λ = 2, ϕ = 0; 5 - 2b/λ = 3,

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It should be noted that the differential radiation pattern of two spaced antennas with the large separation of antennas (2b/λ > 1) has several lug/lobes of identical amplitude (number of lug/lobes it is determined by relation 2b/λ). Curves are given only within the limits of one lug/lobe of radiation pattern with ϕ = 0 (curved 3, 4, 5). For a comparison are plotted/applied also the errors for by the antenna of system with cosinusoidal directional characteristic with $\phi = 0$ and $\phi = \pi$ (curved 1 and 2).

On Fig. 5.7, are represented the maximum errors ($f_{\text{макс}}$) due to the effect of return emitter depending on the separation of antennas $2b/\lambda$ with $k = 0.5$.

Let us calculate the error of the effect of return emitter on the antenna system of radio direction finder with cyclic measurement of phase in high frequency.

Let us suppose that then on the antenna, rotated on radius b (separation of antenna 2b) with frequency Ω, operates ground field E with the angle of the slope of a front of wave β_1 and the field of return emitter $kEe^{i\psi}$ under angle $\rho - \psi = \varphi$. to ground field in horizontal plane and with the angle of the slope of a front of wave β_2.

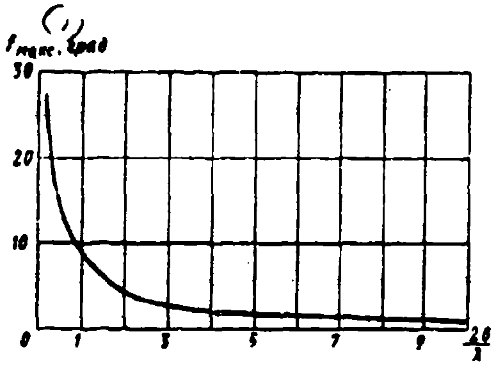

Fig. 5.7. Dependence of maximum errors on the separation by the antenna of system.

Key: (1). deg.

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Let us designate

$$\frac{2\pi b}{\lambda}\cos\beta_1=\delta_1,\ \frac{2\pi b}{\lambda}\cos\beta_2=\delta_2. \tag{5.21}$$

We assume that the reading of the phase of frequency Ω is conducted from the direction of the arrival of ground field E.

Induced in antenna emf will be

$$\dot{E}=Eh\left[e^{j\delta_1\cos\Omega t}+ke^{j\varphi}e^{j\delta_2\cos(\Omega t-\psi_0)}\right]$$

or, after designating $\Omega t = \theta$, we will obtain

$$E = Eh\left\{e^{j\delta_1\cos\theta} + ke^{j[\delta_2\cos(\theta-\psi_0)+\varphi]}\right\}. \qquad (5.22)$$

From expression (5.22) follows that emf in antenna is equal to the sum of two vectors with phase angles $\alpha_1 = \delta_1 \cos\theta$ and $\alpha_2 = \delta_2 \cos(\theta-\psi_0) + \varphi$ and with the amplitude ratio, equal to k (line OA and AB to Fig. 5.8). The total vector OB deviates on phase to angle ζ from vector OA, which corresponds to emf of ground field, moreover

$$\begin{aligned} \operatorname{tg}\zeta \approx \zeta = \frac{k\sin(\alpha_2-\alpha_1)}{1+k\cos(\alpha_2-\alpha_1)} = \\ = k\sin(\alpha_2-\alpha_1)[1-k\cos(\alpha_2-\alpha_1)+\ldots] = \\ = k\sin(\alpha_2-\alpha_1) - \frac{k^2}{2}\sin 2(\alpha_2-\alpha_1)+\ldots \qquad (5.23) \end{aligned}$$

Since

$$\begin{aligned} \alpha_2 - \alpha_1 = \delta_2\cos(\theta-\psi_0) + \varphi - \delta_1\cos\theta = \delta_2\cos\theta\cos\psi_0 + \\ + \delta_2\sin\theta\sin\psi_0 - \delta_1\cos\theta + \varphi = a\sin(\theta-\chi) + \varphi, \end{aligned}$$

where

$$a = \sqrt{\delta_1^2 + \delta_2^2 - 2\delta_1\delta_2\cos\psi_0} \; \overset{\text{and}}{\textcircled{H}} \; \operatorname{tg}\chi = \frac{\delta_1 - \delta_2\cos\psi_0}{\delta_2\sin\psi_0}, \quad (5.24)$$

of (5.23) follows that

$$\begin{aligned} \zeta \approx k\sin[a\sin(\theta-\chi)+\varphi] - \frac{k^2}{2}\sin 2[a\sin(\theta-\chi)+\varphi]+\ldots = \\ = k\{\cos\varphi\sin[a\sin(\theta-\chi)] + \sin\varphi\cos[a\sin(\theta-\chi)]\} + \\ + \frac{k^2}{2}\{\cos 2\varphi\sin[2a\sin(\theta-\chi)] + \sin 2\varphi\cos[2a\sin(\theta-\chi)]\} + \ldots \end{aligned}$$

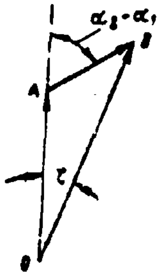

Fig. 5.8. Vector diagram of emf under the effect of the field of reradiation.

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It is decomposed ζ in Fourier-Bessel's series in argument θ and is isolated the term ζ_1 of fundamental frequency θ, since this component affects the measurement of the resulting phase during the determination of the bearing:

$$\zeta_1=2\left[k\cos\varphi J_1(a)-\frac{k^2}{2}\cos 2\varphi J_1(2a)\right]\sin(\theta-\chi)=$$
$$=s\sin(\theta-\chi),$$

where

$$s=2\left[k\cos\varphi J_1(a)-\frac{k^2}{2}\cos 2\varphi J_1(2a)\right].$$

The resulting phase during measurement will be

$$\begin{gathered}
a = a_1 + \zeta_1 = \delta_1 \cos\theta + s\sin(\theta - \chi) = \\
= \delta_1 \cos\theta + s\sin\theta\cos\chi - s\cos\theta\sin\chi = \\
= \sqrt{\delta_1^2 + s^2 - 2\delta_1 s\sin\chi}\,\cos(\theta - f), \\
\operatorname{tg} f \approx \frac{s\cos\chi}{\delta_1 - s\sin\chi}, \qquad (5.25)
\end{gathered}$$

where f - correction to bearing, in absolute value is equal to bearing error.

Formula (5.25) is simplified, if $\delta_1 = \delta_2 = \delta$, i.e., when $\beta_1 = \beta_2 = \beta$.

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In this case, $a = 2\delta\sin\frac{\psi_0}{2}$, $\operatorname{tg}\chi = \operatorname{tg}\frac{\psi_0}{2}$ and $\chi = \frac{\psi_0}{2}$.

$$\operatorname{tg} f = \frac{2\left[k\cos\varphi J_1\left(2\delta\sin\frac{\psi_0}{2}\right) - \frac{k^2}{2}\cos 2\varphi J_1\left(4\delta\sin\frac{\psi_0}{2}\right)\right]\cos\frac{\psi_0}{2}}{\delta - 2\left[J_1\left(2\delta\sin\frac{\psi_0}{2}\right)k\cos\varphi - \frac{k^2}{2}J_1\left(4\delta\sin\frac{\psi_0}{2}\right)\cos 2\varphi\right]\sin\frac{\psi_0}{2}}. \qquad (5.26)$$

When k << 1, it is possible to be restricted to the first terms in numerator and denominator (5.26). Then value f will be

$$f \approx \operatorname{tg} f = -\frac{2k\cos\varphi\cos\frac{\psi_0}{2}\, J_1\left(2\delta\sin\frac{\psi_0}{2}\right)}{\delta}. \qquad (5.27)$$

This expression is analogous (with 5.20). After substituting in (5.27) values ψ_0 and δ, we will obtain

$$f \approx \frac{2k\cos\varphi\cos\left(\frac{\rho-\psi}{2}\right) J_1\left[\ldots\beta\sin\left(\frac{\rho-\psi}{2}\right)\right]}{\frac{2\pi}{\lambda}\, b\cos\beta}. \qquad (5.27')$$

With a small radius of gyration of antenna b (low value 2b) $\delta \to 0$ and $\lim_{\delta\to 0} f = k\cos\varphi\sin(\rho-\psi)$. Result is analogous to formula (5.12).

On Fig. 5.9, is depicted the dependence of the absolute value of the maximum error f of system from separation (radius of a circle of the rotation of antenna). It is accepted that $k = 0.5$, $\beta = 0$, $\psi_0 = \frac{\pi}{2}$.

In phase radio direction finder of two motionless antennas, the bearing error, as this follows from (5.22), is designed from the formula

$$f = \frac{1}{2\delta_1}\operatorname{arc\,tg}\frac{2k\cos\varphi\sin(\delta_1\sin\theta-\delta_2\sin\psi)+}{1+2k\cos\varphi\cos(\delta_1\sin\theta-\delta_2\sin\psi)+} \to$$

$$\to \frac{+k^2\sin[2(\delta_1\sin\theta-\delta_2\sin\psi)]}{+k^2\cos[2(\delta_1\sin\theta-\delta_2\sin\psi)]},$$

where θ is an angle of direction in radio station with perpendicular to lines, that connects antennas.

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When $\theta = 0$, $\delta_1 = \delta_2 = \delta = 2\pi/\lambda\ b \cos \beta$, then

$$|f| = \frac{1}{2\delta} \operatorname{arc\,tg} \frac{2k \cos \varphi \sin(\delta \sin \psi_0) + k^2 \sin(2\delta \sin \psi_0)}{1 + 2k \cos \varphi \cos(\delta \sin \psi_0) + k^2 \cos(2\delta \sin \psi_0)}$$

If in this case $\psi = \frac{\pi}{2}$ and $\varphi = \pi$, then we will obtain the value of the maximum deviation

$$|f|_{\text{макс}} = \frac{1}{2\delta} \operatorname{arc\,tg} \frac{-2k \sin \delta + k^2 \sin 2\delta}{1 - 2k \cos \delta + k^2 \cos^2 \delta}$$

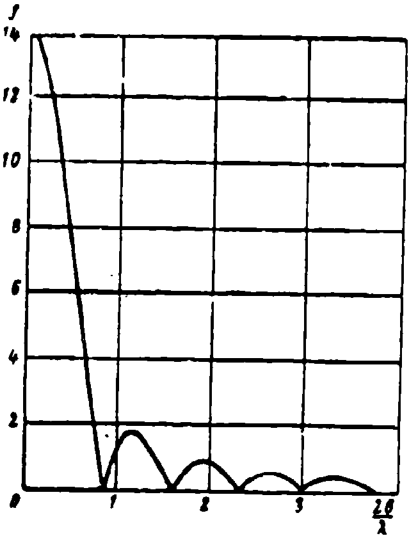

Fig. 5.9. Dependence of the maximum errors of system with the cyclic measurement of phase from the separation of system (with k = 0.5).

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At the low values of k and θ = 0 error function of phase radio direction finder of two motionless antennas approximately corresponds to system of two spaced antennas of amplitude radio direction finder (5.19).

Under actual conditions for the antenna system of radio direction finder, operate the return emitters, arrange/located in different directions also at different distances from by the antenna of system.

Susceptibility by the antenna of the system of radio direction finder to the effect of return emitters it is possible to characterize by mean square error σ_M or idle time by mean error $\tilde{\Delta}$ from the effect of the return emitters, evenly arrange/located around by the antenna of system.

It is assumed that of return emitters the field E_{0M} coincides in the phase with ground field E in center as the antenna of system and has the constant intensity/strength, which is characterized by relation $k=\frac{E_{0M}}{E}=0,1.$ If the return emitter, arrange/located at an angle ψ_0 to the direction of ground field, creates error Δ, then

$$\overline{\Delta^2}=\sigma_M^2=\frac{1}{2\pi}\int_0^{2\pi}\Delta^2 d\psi_0, \overset{\text{and}}{\text{и}}\ \overline{\Delta}=\frac{1}{2\pi}\int_0^{2\pi}|\Delta|\,d\psi_0. \qquad (5.28)$$

Let us calculate σ_M for by the antenna of the system, equivalent to two spaced antennas with separation 2b at whose reading of bearing is realized on the minimum of directional characteristic. We will use for calculation Δ by formula (5.20). Let us designate $2\pi/\lambda = m$. Then with $\phi = 0$

$$\sigma_{м}^{2}=\frac{1}{2\pi}\int\limits_{0}^{2\pi}\left(\frac{k}{mb}\right)^{2}\sin^{2}(mb\sin\psi_{0})\,d\psi_{0}=$$
$$=\frac{1}{2}\left(\frac{k}{mb}\right)^{2}[1-J_{0}(2mb)]. \qquad (5.29)$$

With k = 0.1

$$\sigma_{м}=\left[\frac{0{,}1\cdot 57{,}3}{\sqrt{2}\,mb}\sqrt{1-J_{0}(2mb)}\right]=\frac{4{,}05}{mb}\sqrt{1-J_{0}(2mb)},\quad \text{deg.}$$

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For the antenna system with cosinusoidal directional characteristic, which has

$$\frac{2\pi b}{\lambda}\ll 1 \quad \text{and} \quad J_{0}(2mb)\approx 1-\left(\frac{2mb}{2}\right)^{2},$$
$$\sigma_{мк}=4{,}05^{\circ}.$$

Relation of the mean square errors of the larger-base and cosinusoidal systems

$$\frac{\sigma_{м}}{\sigma_{мк}}=\frac{\sqrt{1-J_{0}(2mb)}}{mb}. \qquad (5.30)$$

On Fig. 5.10, are given to the dependence $\frac{\sigma_{м}}{\sigma_{мк}}$ on the separation of antennas for a radio direction finder of two spaced antennas with the reading of bearing on the minimum (curved 1), and also of two diverse coaxial framework (curved 2).

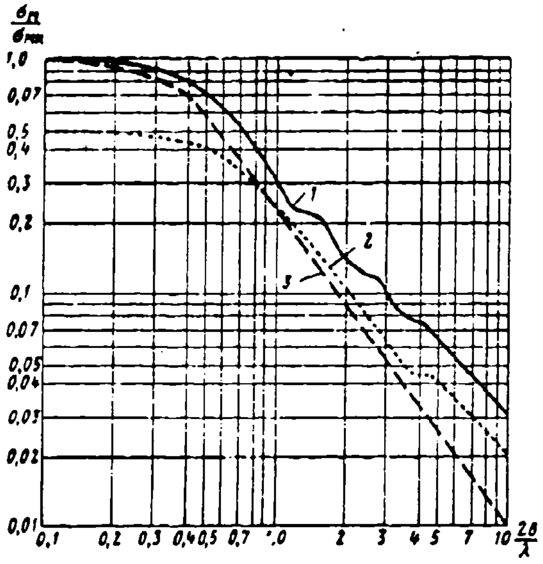

Fig. 5.10. Dependence of the root mean square values of local errors from the separation of the antennas: 1 - for two spaced antennas; 2 - for two diverse framework; 3 - for a system with the cyclic measurement of phase.

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From Fig. 5.10, it is evident that with an increase in the separation of antennas to 2b = (3-4) λ the mean square error $\sigma_{м}$ decreases 10 - 15 times in comparison with error $\sigma_{мк}$ for by the antenna of system with the cosinusoidal characteristic of directivity. With a further increase in the separation σ_M it decreases more slowly and with 2b = 10 λ it decreases 2 more times.

If one considers that the large minor lobes, which there are at the radiation pattern of two spaced antennas when 2b/λ > 1, in reality are suppressed in system with acute/sharp directional characteristic, then in system with large separation a decrease in the errors due to return emitters must be larger than in Fig. 5.10.

For a radio direction finder with the cyclic measurement of phase in high frequency, the calculation of mean square error from the effect of the return emitters, evenly arrange/located around by the antenna of system, is difficult. For calculation ille time of mean error we will use expression (5.27)

$$\bar{\Delta}_M = \frac{1}{2\pi}\frac{2k\cos\varphi}{\delta}\int_0^{\pi}\left|\cos\frac{\psi_0}{2}J_1\left(2\delta\sin\frac{\psi_0}{2}\right)\right|d\psi_0. \quad (5.31)$$

Let us replace of variable $2\delta\sin\frac{\psi_0}{2}=x$:

$$\bar{\Delta}_M = \frac{2}{\pi}\frac{k\cos\varphi}{\delta^2}\int_0^{2\delta}|J_1(x)|\,dx = \frac{2k\cos\varphi}{\pi\delta^2}\Big\{[1-|J_0(2\delta)|]+$$

$$+2\sum_{k=1}^{n}(-1)^k J_0(2\delta_k)\Big\}. \quad (5.32)$$

Since $J_1(x)$ - sign-changing function, integration is conducted between values δ_n, by corresponding n to zero $J_1(x)$. General solution is equal to the sum of the indicated particular integrals. With small δ

$$1-J_0(2\delta)=\delta^2$$

and

$$\overline{\Delta}_{\text{м}}=\frac{2}{\pi}k\cos\varphi, \qquad (5.32')$$

which coincides with formula for two spaced antennas with a small separation of antennas.

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To Fig. 5.10 (curved 3) are plotted/applied also obtained by numerical integration of the relation of the mean square errors of systems with the cyclic measurement of the phase in high frequency and cosinusoidal depending on separation 2b/λ [5.12].

It is possible to obtain expression for an error $\sigma_{\text{м}}$ for a radio direction finder with the reading of bearing from maxima, if we, for example, to assume that the radiation pattern of antenna system has one lug/lobe with a width of $B_{0,5}$ during a decrease in the power of signal 2 times [5.12].

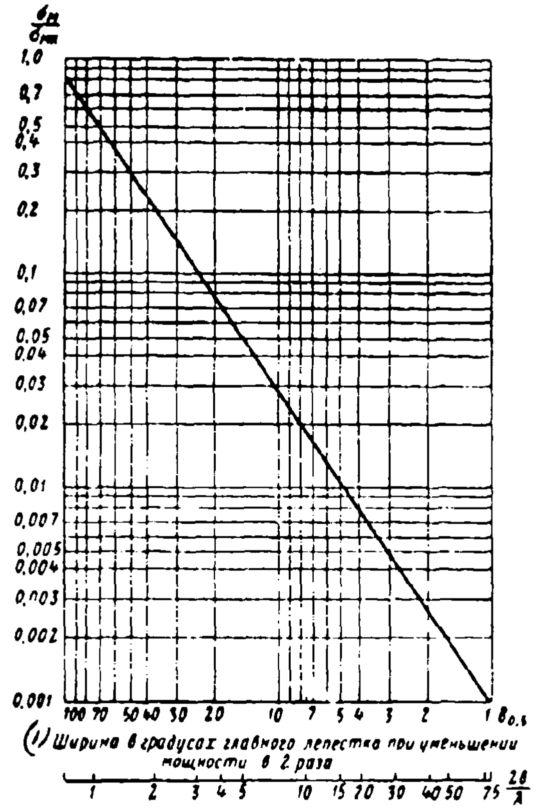

Fig. 5.11. dependence of the root mean square values of local errors by the single-lobe antenna of system from the width of lug/lobe and separation of antennas.

Key: (1). Width in the degrees of major lobe during a decrease in the power 2 times.

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Within the limits of lug/lobe, accepted expression for directional characteristic

$$F(\theta)=\sin\left(\frac{\pi\theta}{2B_{0,5}}\right),$$

of the outside limits of lug/lobe of characteristic are assumed zero values.

Mean square error will be

$$\sigma_{м}=1,45B_{0,5}^{1,5}, \text{ deg}$$

where $B_{0,5}$ is expressed in radians.

To Fig. 5.11, is given the dependence $\frac{\sigma_{м}}{\sigma_{мп}}$ on the width of lug/lobe $B_{0.5}$. On the axis of abscissas, are plotted/applied also the approximate values of ratio $2b/\lambda$, designed from the assumption that the lug/lobe of directional characteristic is created by two spaced antennas.

5.4. Types of return emitters.

In §§ 5.5 and 5.6 it is examined the effect of the return emitters of two types:

- analogous to antennas, i.e., the elongated emitters in which one size/dimension many times is be greater the size/dimensions of section;

- in the form of duct-framework from wires (locked duct, open circuit).

Inverse-exciter can it is located on very close distance from by the antenna of the system of radio direction finder, for example under conditions of ship and aircraft, and at large distances - under conditions of the ground-based installation of radio direction finder. The method of the examination of antennas is different for both cases. In the first case it is necessary to proceed from the near field of antenna taking into account induction field. In the second case affects only the radiation field.

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The action of inversely-radiating circuit (framework) can be observed aboard ship and on aircraft, and also in ground-based radio direction finder. In § 5.6 it is examined the effect of duct in ship radio direction finder. It is assumed, that the size/dimensions of duct are less than the wavelength. If the size/dimensions of duct are commensurable with wavelength and it is more than it, then the effect of this duct can be considered as cumulative effect of its sides [5.10].

To account for the effect of the conducting object/subjects, all three size/dimensions of which of one order are commensurable with wavelength (metallic installations, mountains, buildings), one ought not to have solved the problems of wave diffraction for the different cases of practice. These questions are insufficiently developed. In § 5.7 is given only one such case - application/use of theory of diffraction on the calculation of the deviation, caused by the metallic body of ship and aircraft.

5.5. Action of the antenna, located near radio direction finder.

Antenna is arrange/located in immediate proximity of radio direction finder.

To the framework of radio direction finder at an angle of p the center-line plane of ship, approaches the electromagnetic wave with the intensity/strength of electric field E and with normal polarization.

At certain distance from the framework, is arrange/located the vertical inverse-exiter A. The angle between the direction "framework-antenna" and the center-line plane of ship let us designate by ψ (Fig. 5.5).

Earlier we were obtained for the deviation, caused by return emitter, expressions (5.14) in the case when there is compensation for antenna effects [1], and (5.11) in the case when the compensation for antenna effects is absent.

FOOTNOTE [1]. The effect of compensative antenna (§ 7.12) it is not considered. ENDFOOTNOTE.

For inverse-exiter, which is the nondirectional system.

$$F(\theta_0,\ \psi)=F(\theta_0,\ \psi_0)=1,\ h_{\text{ои}}=h_{\text{а}},\ z_{\text{ои}}=z_{\text{а}},\ \varphi_{z0}=0.$$
$$\varphi=\varphi_1+\varphi_{\text{а}0}+\frac{2\pi}{\lambda}d(1-\cos\psi_0).$$

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From (5.3) it follows that

$$k = \frac{ah_{\text{д}}}{z_{\text{д}}} = \text{const}$$

independent of p and 9. Therefore formulas (5.11) and (5.14) retain their form without any changes.

From (5.14) follows that

$$\sin f = k \cos \varphi \cos \psi \sin q - k \cos \varphi \sin \psi \cos q.$$

Let ø does not depend on q.

For small deviations we replace sin f on f:

$$f = B \sin q + C \cos q, \qquad (5.33)$$

where $B = k \cos \varphi \cos \psi, \; C = -k \cos \varphi \sin \psi.$

Thus when in radio direction finder is compensation for antenna effects or the field of antenna coincides in phase with ground field, the vertical inverse-exiter creates the semicircular deviation with coefficients B and C.

The signs of the coefficients depend on ψ, i.e. on the location of antenna relative to the framework.

The signs of coefficients B and C are represented on Fig. 5.12.

When in radio direction finder there is no equipment/device for the compensation for antenna effects, deviation is determined by formula (5.11).

The representation of how changes in this case deviation depending on the character of antenna resistance, gives Fig. 5.13a. On it is depicted dependence of f on p for different values ϕ_1 from 0 to 180° when $k=0,5$, $\psi=0$, $\varphi_2+\varphi_3=0$ (compensation for antenna effects it is absent). On Fig. 5.13b, is given dependence of f on p for $k=0,5$, $\psi=0$, when distance $d = 0.5\lambda$ and ϕ_1 it varies from -180 to +180°, also when the compensation for the antenna of effects. On figure are plotted/applied inclined straight lines, $q = p - f$, so that is visible dependence of f on q.

If we examine dependence of f on q, then sometimes from curved (Fig. 5.13b) for the determined values of q and ϕ_1 is obtained two values f.

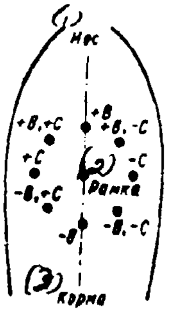

Fig. 5.12. Signs of the coefficients B and C with the different locations of inverse-exiter.

Key: (1). Nose. (2). Framework. (3). Forages.

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This frequently is encountered in practice when is determined deviation f depending on q. Multiformity f disappears, if we examine dependence of f on p.

From Fig. 5.13a, it follows that if we disregard increment for a field from inverse-exiter (case of medium-frequency waves or very close antenna location to by antenna to the system of radio direction finder), then with small detunings of antenna, when the reactance of antenna it is small and $\phi_1 = 0$, deviation it reaches maximum values (to 30°) with q = 90° and q =270°, i.e., has semi-circular law from q.

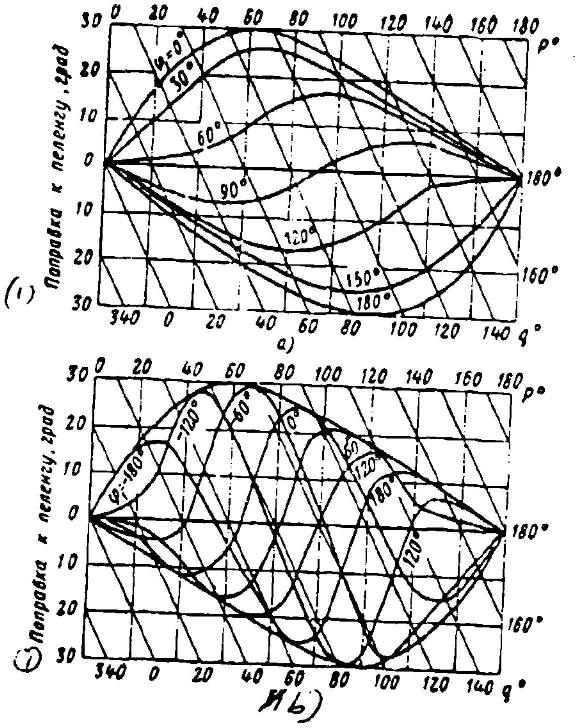

Fig. 5.13. Error, caused by inverse-exciter when the compensation for the antenna effects is absent,: a) k = 0.5; $\psi=0; \Phi_2 + \Phi_3 = 0$; b) k = 0.5; $\psi=0; \Phi_2 + \Phi_3 = \pi(1 - \cos p)$.

Key: (1). Correction to bearing, deg.

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With an increase in the detuning of the inverse-exciter the law governing deviation changes for large mismatches of antenna, when $\phi_1 \approx 90^\circ$, deviation has maximum a value $f_{\text{макс}}=8^\circ$ with q = 45, 135, 225 and 315°. the law governing deviations becomes one fourth from q.

From Fig. 5.13b, it follows that if we consider increment for the field of reradiation, then deviation reaches maximum 30° with certain detuning of the inverse-exciter. The laws governing deviation for case Fig. 5.13b are obtained more complex than for case Fig. 5.13a.

In order to judge the effect of the inverse-exciter it is necessary to know, as they change with k and ϕ depending on the site of installation of the framework of radio direction finder, and also on the relationship/ratio of the frequency of direction finding and natural frequency of antenna.

Let us examine first as they change with k and ϕ depending on the arrangement/permutation of the framework of radio direction finder relative to the inverse-exciter. For this, it is necessary to find expression for electromagnetic field on close distance from the inverse-exciter. P. A. Ryazin [5.7] designed the field, created by rectilinear antenna of near zone as sum of fields from elementary radiation currents, under the assumption that the current distribution along the antenna is sinusoidal.

In our examination we assume that the reactance of antenna is equal to zero. This corresponds to such frequency at which along the length l_a of the grounded antenna is placed the odd number of quarter-wave lengths i.e. $l_a = \frac{2n-1}{4}\lambda$, where n any whole.

Expression for the which interests us in the case of reception/procedure to the framework magnetic intensity from the vertical wire antenna by length $l_a = \frac{2n-1}{4}\lambda$ obtained in [5.7] in the following form:

$$\dot{H}_{0н} = \frac{120 l_a E}{(2n-1)\pi R_a r} \cos\left[\frac{m}{2}(r_в - r_н)\right] e^{j(\varphi_з + \varphi_з)}, \qquad (5.34)$$

moreover $\phi_{20} = 0$, $\phi_з - \phi_{30} = m/2\ (r_в + r_н)$, ϕ_{30} depends on n, where r is
r -
a distance from the framework to antenna on perpendicular to antenna;

$r_в$ - a distance from the framework to the top of antenna;

$r_н$ - a distance from the framework to the image of the top of antenna:

R_a - the effective resistance of antenna.

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The amplitude of field component, cophasal with the field of transmitter, is determined by the expression

$$H'_{\text{он}}=\frac{120l_{\text{а}}E}{(2n-1)\pi R_{\text{а}}r}\cos\left[\frac{m}{2}(r_{\text{в}}-r_{\text{п}})\right]\cos\varphi, \quad (5.35)$$

where

$$\varphi=\varphi_{\text{з}}+\varphi_{\text{зо}}-\frac{m}{2}(r_{\text{в}}+r_{\text{п}});$$

$$\varphi_{\text{з}}=\frac{2\pi}{\lambda}r\cos\psi_{\text{о}}. \qquad (5.36)$$

Let us examine two cases, that are of greatest practical interest.

1. Along the length of antenna, is placed quarter wavelength, i.e., $l_{\text{а}}=\frac{\lambda}{4}$. For this antenna

$$R_{\text{а}}=36{,}6\ \Omega,\ m=\frac{2\pi}{\lambda}=\frac{\pi}{2l_{\text{а}}}.$$

From formulas (5.35) and (5.36) follows that

$$H'_{\text{он}}=1{,}04E\frac{l_{\text{а}}}{r}\cos\left[\frac{\pi}{2}\left(\frac{r_{\text{в}}-r_{\text{п}}}{2l_{\text{а}}}\right)\right]\cos\varphi,$$

whence

$$k=\frac{H_{\text{он}}}{E}=1{,}04\frac{l_{\text{а}}}{r}\cos\left[\frac{\pi}{2}\left(\frac{r_{\text{в}}-r_{\text{п}}}{2l_{\text{а}}}\right)\right]. \qquad (5.37)$$

in this case

$$\varphi_{\text{зо}}=\frac{\pi}{2} \text{ и } \varphi=\varphi_{\text{з}}+\frac{\pi}{2}-\frac{\pi}{2}\left(\frac{r_{\text{в}}+r_{\text{н}}}{2l_{\text{а}}}\right). \qquad (5.38)$$

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2. Along the length of antenna, are placed three quarter wavelengths, i.e., $l_{\text{а}}=\frac{3}{4}\lambda$. In this case $R_{\text{а}}=55.2$ ohm, $m=\frac{3\pi}{2l_{\text{а}}}$. From formulas (5.35) and (5.36) follows that

$$H'_{\text{он}}=0{,}230\mathrm{E}\frac{l_{\text{а}}}{r}\cos\left[\frac{3\pi}{2}\left(\frac{r_{\text{в}}-r_{\text{н}}}{2l_{\text{а}}}\right)\right]\cos\varphi,$$

whence

$$\mathrm{k}=\frac{H_{\text{он}}}{\mathrm{E}}=0{,}230\frac{l_{\text{а}}}{r}\cos\left[\frac{3\pi}{2}\left(\frac{r_{\text{в}}-r_{\text{н}}}{2l_{\text{а}}}\right)\right], \qquad (5.39)$$

where

$$\varphi=\varphi_{\text{з}}+\frac{\pi}{2}\left(1-3\frac{r_{\text{в}}+r_{\text{н}}}{2l_{\text{а}}}\right), \quad \text{если } \frac{r_{\text{в}}-r_{\text{н}}}{2l_{\text{а}}}<\frac{1}{3},$$

$$\varphi=\varphi_{\text{з}}+\frac{3\pi}{2}\left(1-\frac{r_{\text{в}}+r_{\text{н}}}{2l_{\text{а}}}\right), \quad \text{если } \frac{r_{\text{в}}-r_{\text{н}}}{2l_{\text{а}}}>\frac{1}{3}.$$

[если = if]

The maximum values k for the different locations of the framework of radio direction finder relative to the inverse-exciter are depicted on Fig. 5.14a, when $l_a = \frac{1}{4}\lambda$, and on Fig. 5.14b when $l_a = \frac{3}{4}\lambda$.

Fig. 5.14, are shown also for the different places of arrangement/permutation by the antenna of the system of the value of maximum deviation $f_{макс}$ = arc sin k when $\phi = 0$, when $\phi \neq 0$ value of deviation they will be smaller. With $\phi = (\pi/2)$ f = 0 and $\left(\frac{A}{B}\right)_{макс} = k$. These curves it is possible to use during the site of installation of radio direction finder aboard ship. From curves it follows that the deviation decreases with the lift of the framework above the hull of ship. For a decrease in the effect of the mast of ship, it is expedient to assemble the framework on the top of mast on the line of the symmetry of mast. As this follows from the curves Fig. 5.14, for equidistances between the framework and the antenna when $l_a = \frac{3}{4}\lambda$ of error, it is less than when $l_a = \frac{1}{4}\lambda$.

Let us determine the effect of the detuning of the natural frequency of antenna relative to the frequency of direction finding ω on an example of the short antenna which is tuned to a frequency ω_a by supplementary reactive cell/elements.

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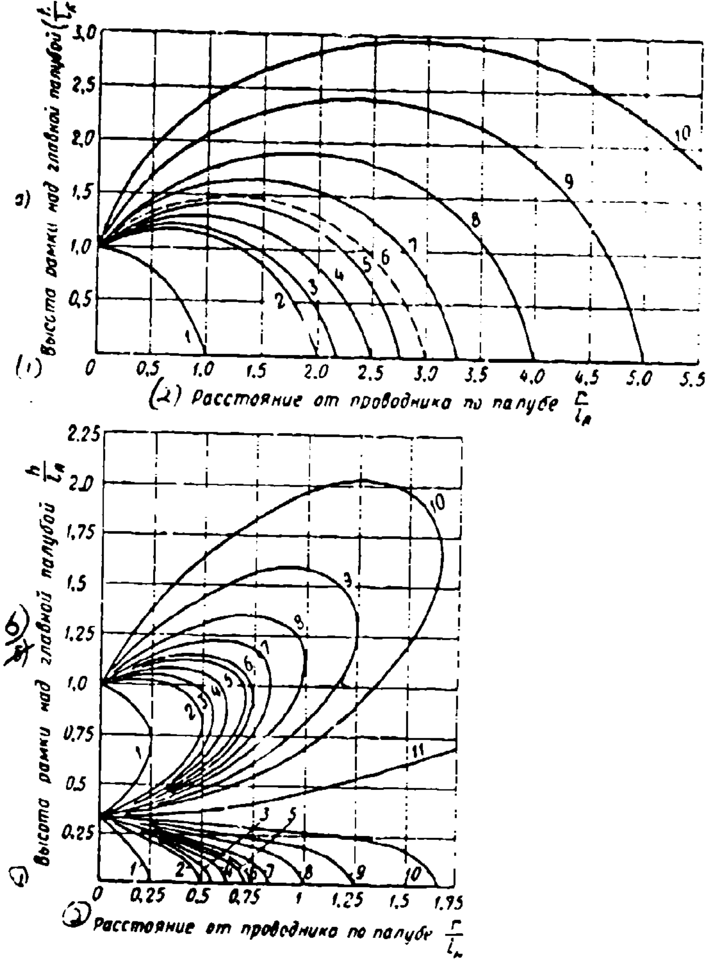

Fig. 5.14. Effect of the secondary field of the grounded wire in the vertical plane: a - with $l_в=1.0$; b - when $l_в=3.0$.

1) $k = 1, \quad l_{макс} \leq 90°;$
2) $k = 0.5, \quad l_{макс} \leq 30°;$
3) $k = 0.45, \quad l_{макс} \leq 26.7°;$
4) $k = 0.4, \quad l_{макс} \leq 23.6°;$
5) $k = 0.35, \quad l_{макс} \leq 20.5°;$
6) $k = 0.333, \quad l_{макс} \leq 19.5°;$
7) $k = 0.3, \quad l_{макс} \leq 17.5°;$
8) $k = 0.25, \quad l_{макс} \leq 14.5°;$
9) $k = 0.2, \quad l_{макс} \leq 11.5°;$
10) $k = 0.15, \quad l_{макс} \leq 8.6°;$
11) $k = 0, \quad l_{макс} = 0.$

Key: (1). Height/altitude of the framework above the main deck (2). Distance from conductor on deck

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In this case, $\Phi_c + \varphi_3 = 0$, i.e., the radio direction finder works on average/mean and long waves.

Let us designate:

$Z_a = R_a + jX_a$ - the complete active and reactance of aatenna circuit;

$z_a = \sqrt{R_a^2 + X_a^2}$ - modulus of resistance Z_a.

On the basis (5.3), for the inverse-exciter

$$\frac{E_{ов}}{E} = \frac{ah_a}{Z_a} = \frac{ah_a}{z_a^2} R_a - j\frac{ah_a}{z_a^2} X_a = k' + jk'',$$

where a is the coefficient of proportionality.

If one assumes that h_a does not change with frequency, then dependences k' and k" on $\frac{\omega_a}{\omega}$ will coincide with dependences for antenna circuit $\frac{R_a}{z_a^2}$ and $\frac{X_a}{z_a^2}$ on relation $\frac{\omega_a}{\omega} = \frac{\lambda}{\lambda_a}$. These dependences are represented on Fig 5.15. The sharpness of curves is determined by the quality of antenna circuit. From Fig. 5.15, it follows that the greatest values of the coefficients of semicircular deviation and diffuseness of the minimum are obtained, when the natural frequency of antenna is close to the frequency of direction finding.

2. Antenna is arrange/located at a great distance from radio direction finder.

Let us examine the action of the vertical grounded antenna, arrange/located at a great distance from by the antenna of system radio direction finder.

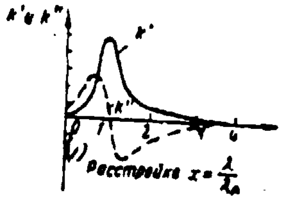

Fig. 5.15. Dependence of the field of reradiation on the detuning of antenna.

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Let the vertical radiator have effective height h_a and is located at a distance r from the framework of direction finder. The strength of the field of the oriented transmitter let us designate E.

In antenna it is induced by emf

$$E_a = Eh_a,$$

where $h_a = \frac{\lambda \operatorname{tg}\left(\frac{\pi l_a}{\lambda}\right)}{2\pi}$ is the effective height of antenna.

The quite high field of reradiation will be, when $l_a = \frac{\lambda}{4}$. In this case

$$h_a = \frac{\lambda}{2\pi},\quad E_a = E\frac{\lambda}{2\pi},\quad R_a = 36{,}6 \text{ ом.}\qquad \text{ohm}$$

The field of reradiation, created by the distant antenna, will be

$$E_{он} = \frac{377 I_a h_a}{\lambda r} = \frac{377 E \frac{\lambda}{2\pi} \frac{\lambda}{2\pi}}{\lambda r 36{,}6} \approx 0{,}25 \frac{\lambda}{r} E$$

or, since $\lambda = 4 l_a$,

$$E_{он} = \frac{l_a}{r} E \text{ and } k = \frac{E_{он}}{E} = \frac{l_a}{r}.$$

The greatest action of antenna will be, when the phase of the field of reradiation in location by the antenna of the system of radio direction finder coincides with the phase of ground field (transmitter) or differs from it in phase to 180°; in this case maximum deviation is determined by the formula

$$f_{макс} = k = \frac{l_a}{r} \text{ rad}, \tag{5.40}$$

On the basis of the permissible deviation into 1°, from (5.40) we have

$$f_{макс} < \frac{1}{57},\quad \frac{l_a}{r} < \frac{1}{57},\quad \text{or } r > \sim 50 l_a.$$

Thus, so that the deviation from the vertical quarter-wave grounded antenna does not exceed 1°, the distance of the vertical wire antenna from the framework of radio direction finder must be 50 (and more) times of the more height of the vertical wire antenna.

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The maximum deviation, produced by the effect of the horizontal wire, suspend/hung from height/altitude h at a distance r from by the antenna of the system of radio direction finder, is determined from the formula

$$f \approx \frac{\lambda h}{r^2}.$$

So that the error does not exceed 1°, it is necessary to fulfill requirement $\frac{\lambda h}{r^2} \leqslant 0.02$ or $r \geqslant 7\sqrt{\lambda h}$.

Thus, for instance, for a wave $\lambda = 100$ m at the height/altitude of the supports of the suspension of wire $h = 10$ m for the limitation of error by value of 1° must be made condition $r \geqslant 7\sqrt{1000}$ or $r \geqslant$ 250 m.

5.6. Action of inversely-radiating framework.

The inversely-radiating framework frequently calls loop. The framework of radio direction finder with size/dimensions, much smaller than the wavelength, it is located within the inversely-radiating framework (Fig. 5.16). For determining the effect of the inversely-radiating framework, we proceed from (5.3), where h_{ou} is the effective height of the inversely-radiating framework. In the general case the action of the framework of the equivalently two spaced antennas.

If the size/dimensions of the inversely-radiating framework are small in comparison with wavelength and the direction-finding framework is arrange/located in the center of the inversely-radiating framework, then

$$F(\theta_0, \psi_0) = \cos(p - \psi) = \cos\psi_0, \quad aF(\theta_0, \psi) = \frac{\omega M_m}{h_p},$$

where M_m is maximum mutual inductance between the direction-finding and inversely-radiating framework;

h_p - the effective height of the direction-finding framework.

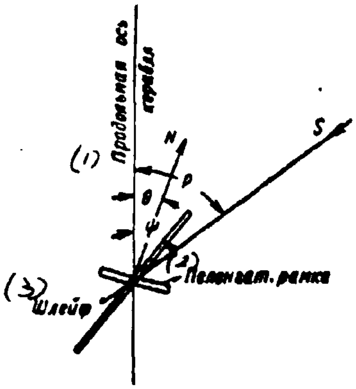

Fig. 5.16. Framework as return emitter.

Key: (1). The longitudinal axis of the ship. (2). Direction finder is the framework. (3). Loop.

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From (5.3) and (5.4) we have

$$k=\frac{\omega M_m h_{\text{ов}}}{z_{\text{ов}} h_{\text{р}}}\cos\psi_0=m\cos\psi_0,\quad \varphi=\varphi_1\pm\frac{\pi}{2},\qquad (5.41)$$

where

$$m=\frac{\omega M_m h_{\text{ов}}}{z_{\text{ов}} h_{\text{р}}}.$$

In the case in question the resistor/resistance of inversely-radiating framework must be predominantly either inductive or capacitive, i.e.,

$$Z_{\text{ои}} = \pm jX_{\text{ои}}.$$

Therefore field $E_{\text{ои}}$ coincides in phase with field E of transmitter.

Utilizing expressions (5.14)and (5.41), and also taking into account that with the reading of bearing 0 = q, p = q + f, we obtain the expression, which connects bearing deviation f with q and ψ;

$$\sin f = m\cos(q-\psi+f)\sin(q-\psi),$$

whence

$$\operatorname{tg} f = \frac{m\sin 2(q-\psi)}{(2+m)-m\cos 2(q-\psi)} = \frac{\frac{m}{2+m}\sin 2(q-\psi)}{1-\frac{m}{2+m}\cos 2(q-\psi)}. \quad (5.42)$$

Let us replace $q-\psi=q_1$, i.e. let us select as the reference point of bearing not longitudinal axis of ship, but the projection of the inversely-radiating framework on the body of ship or aircraft.

Then

$$\operatorname{tg} f = \frac{\frac{m}{2+m}\sin 2q_1}{1-\frac{m}{2+m}\cos 2q_1}.$$

Let us designate $\frac{m}{2+m} = \mathrm{D}$. (5.43) whence

$$\operatorname{tg} f=\frac{D\sin 2q_1}{1-D\cos 2q_1}. \qquad (5.44)$$

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Expanding equation (5.44) in Fourier series, we have for the low values of f

$$f=D\sin 2q_1+D^2\sin 2q_1\cos 2q_1+\ldots= \\ =D\sin 2q_1+K\sin 4q_1+\ldots, \qquad (5.45)$$

where

$$K=\frac{D^2}{2}. \qquad (5.46)$$

In formulas (5.43) - (5.46) D and K are expressed in radians.

Let us find from (5.45) expression for f at different ψ. If ψ=0, i.e., an inversely-radiating framework is arrange/located along the longitudinal axis of ship (aircraft), then q_1 = q and equation (5.45) will take the form:

$$f=D\sin 2q+K\sin 4q+\ldots \qquad (5.45')$$

Thus, an inversely-radiating framework, arrange/located along the longitudinal axis of ship or aircraft, in the plane, which contains the axis of the direction-finding framework, creates quadrantal and octant deviation.

When $\psi = 90°$, i.e., an inversely-radiating framework is arrange/located perpendicularly to the longitudinal axis of ship and its plane contains the axis of the direction-finding framework, from (5.45) we will obtain:

$$f = D \sin(2\varphi + 180°) + K \sin(4\varphi + 360°) + \ldots = \\ = -D \sin 2\varphi + K \sin 4\varphi \ldots \qquad (5.47)$$

When ψ is equal to any angle, then

$$f = D \sin(2\varphi + 2\psi) + K \sin(4\varphi + 4\psi) = D \cos 2\psi \sin 2\varphi + \\ + D \sin 2\psi \cos 2\varphi + K \cos 4\psi \sin 4\varphi + K \sin 4\psi \cos 4\varphi = \\ = D_1 \sin 2\varphi + E \cos 2\varphi + K_1 \sin 4\varphi + L \cos 4\varphi + \ldots, \qquad (5.48)$$

where

$$D_1 = D \cos 2\psi; \quad K_1 = K \cos 4\psi; \quad E = D \sin 2\psi, \quad L = K \sin 4\psi.$$

Thus, an inversely-radiating framework at any angle ψ to the longitudinal axis of ship (aircraft) with the plane, passing through the axis of the direction-finding framework, creates the deviation of form (5.48), i.e., to one fourth and octant law.

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It is possible to show that an inversely-radiating framework, displaced with respect to the rotational axis of the direction-finding framework, produces the deviation of form (5.48) and, furthermore, constant the depending on wavelength component of deviation (coefficient A) [1.6].

From formulas (5.41) and (5.42) it is possible to determine the law of a change in the deviation with frequency. For this, it is necessary to determine a change in the resistor/resistance of an inversely-radiating framework with frequency.

An inversely-radiating framework can be locked and that which was extended. In its first case they call the inductive or locked duct, in the second case - by the capacitive or open circuit.

The resistor/resistance of inductive duct for the frequencies smaller than its own, it is inductive. Of capacitive duct for the frequencies smaller than its own, the resistor/resistance of duct has capacitive character, for the frequencies, greater than its own, inductive character.

The sign of the coefficients of deviation depends on that, is arrange/located the direction-finding framework within duct or outside it.

From the analysis of formulas (5.41) and (5.42) it is possible to make following conclusions.

1. At natural frequency of inversely-radiating framework, and also at other frequencies when $X_{ou}=0$, inversely-radiating framework does not create deviation, but is produced diffuseness of bearing.

2. Of capacitive duct at frequencies smaller than its own, for direction-finding framework within duct, coefficients of deviation D and E are positive and with decrease in frequency decrease, striving for certain limit.

For the frequencies of the direction finding, large their own, and for the same conditions coefficients D and E are negative and with an increase in the frequency decrease, striving for zero.

3. Of inductive duct at frequencies smaller than its own, and for direction-finding framework within duct coefficients D and E have minus sign and with decrease in frequency decrease, striving for constant limit.

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4. Coefficients of quadrantal deviation D and E for waves of direction finding, close to their own wave of duct, have two maximums (positive and negative).

5) For after becoming the arrangement/permutations of the direction-finding framework of outside the field inversely-radiating circuit the signs of deviation, stipulated in p. 2 and 3, they change by reverse/inverse.

Table 5.1 gives the signs of the coefficients of the fourth deviation D and E for the frequencies of the direction finding smaller than the natural frequency inversely-radiating circuit.

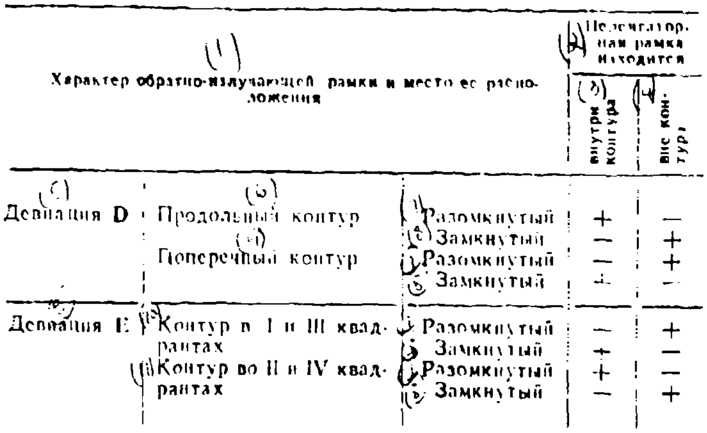
For the frequencies of the direction finding, large of the natural frequency of duct, the coefficients of deviation D and E have opposite signs.

5.7. Deviation caused by the hull of ship.

In work [5.9] is found the expression for a secondary field in any point under the influence of external electromagnetic sex/floor on the ideally conducting semicylinder of infinite length, which lies on the ideally conducting surface.

We will pause on the case of radiowave propagation in horizontal plane ($\beta = 0$).

Table 5.1. Signs of the coefficients of the quadrantal deviation D and E.



(1) Характер обратно-излучающей рамки и место ее расположения			(2) Пеленгаторная рамка находится	
			(3) внутри контура	(4) вне контура
(5) Девиация D	(6) Продольный контур	(7) Разомкнутый	+	—
		(8) Замкнутый	—	+
	(9) Поперечный контур	(7) Разомкнутый	—	+
		(8) Замкнутый	+	—
(10) Девиация E	(10) Контур в I и III квадрантах	(7) Разомкнутый	—	+
		(8) Замкнутый	+	—
	(11) Контур во II и IV квадрантах	(7) Разомкнутый	+	—
		(8) Замкнутый	—	+

Key: (1). Character of inversely-radiating framework and its location. (2). The direction-finding framework is located. (3). within duct. (4). outside duct. (5). Deviation. (6). Longitudinal duct. (7). Extended. (8). Locked. (9). Transverse duct. (10). Duct in I and III quadrants. (11). Duct in II and IV quadrants.

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We will designate: ρ_0 - radius of cylinder; ρ is a distance of observation point from the center of cylinder; H_m - magnetic component of normal-polarized electromagnetic field of transmitter; H_0 H_0 is the field at any point, obtained as a result of diffraction; p the angle between direction of propagation and the axis of semicylinder; q - the angle between the direction of the resulting field and by the axis of semicylinder.

In work is assumed, that $\rho/\lambda \leqslant 1$ and $\rho_0/\lambda \leqslant 1$.

Expressions for field components H_0 above the apex/vertex of semicylinder (Z axis is directed along the axis of semicylinder) it takes the form:

$$\begin{gathered} H_z \approx -2H_m \sin p, \\ H_\varphi = -2H_m\left[1+\left(\frac{\rho_0}{\rho}\right)^2\right]\cos p, \qquad (5.49) \\ H_r = 0. \end{gathered}$$

Expressions for the components H_0 above the plane of the semicylinder:

$$\begin{gathered} H_z = -2H_m \sin p,\ H_\varphi = \pm 2H_m\left[1-\left(\frac{\rho_0}{\rho}\right)^2\right]\cos p, \\ H_r = 0. \qquad (5.50) \end{gathered}$$

Signs (+) and (-) are taken depending on that, where is located observation point - in front of semicylinder or from behind it.

From (5.49) it follows that above the apex/vertex of the semicylinder

$$\mathrm{tg}\, q = \frac{H_z}{H_\varphi} = \frac{1}{1+\left(\frac{\rho_0}{\rho}\right)^2}\,\mathrm{tg}\, p = a\,\mathrm{tg}\, p, \qquad (5.51)$$

where

$$a = \frac{1}{1+\left(\frac{\rho_0}{\rho}\right)^2}. \qquad (5.52)$$

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Dependence (5.51) corresponds to bearing error (4.16). Therefore

$$\mathrm{tg}\, f \approx f = \frac{\frac{1-a}{1+a}\sin 2q}{1-\frac{1-a}{1+a}\cos 2q} = \mathrm{D}\sin 2q + \mathrm{K}\sin 4q + \dots \qquad (5.53)$$

where

$$\mathrm{D} = \frac{1-a}{1+a} \text{ и } \mathrm{K} = \frac{\mathrm{D}^2}{2}. \qquad (5.54)$$

In (5.54) the coefficients of deviation D and K, are expressed in radians.

At the low values of D, the deviation bears purely one fourth character and is expressed

$$f = \mathrm{D}\sin 2q.$$

If we desigaate

$$\frac{\rho_0}{\rho} = m, \qquad (5.55')$$

then, substituting (5.52) in (5.54) and taking into account (5.55), we obtain

$$\mathrm{D} = \frac{m^2}{2+m^2}. \qquad (5.56)$$

Thus, the conducting semicylinder of infinite length in the points, which correspond to its apex/vertex, produces the deviation of the form

$$\mathrm{tg}\, f = \frac{\frac{m^2}{2+m^2}\sin 2q}{1-\frac{m^2}{2+m^2}\cos 2q}. \qquad (5.57)$$

Differentiating equation (5.57) for q, we obtain

$$f_{\text{макс}} = \mathrm{arc\,tg}\,\frac{m^2}{2\sqrt{1+m^2}} \approx \frac{m^2}{2\sqrt{1+m^2}}. \qquad (5.58)$$

The angle $q_{\text{макс}}$, which corresponds to deviation $f_{\text{макс}}$, is expressed as

$$q_{\text{макс}} = \frac{1}{2}\arccos\frac{m^2}{2+m^2}. \qquad (5.59)$$

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Field, created by the metallic hull of ship on long waves, on its action in the first approximation, of the analogous with the field infinitely long semicylinder of which $\rho_0/\lambda \ll 1$ and $\rho/\lambda \ll 1$. (values of $\rho/\lambda = 0.3$ and $\rho_0/\lambda = 0.3$ in given formulas are valid with an accuracy to 5o/o). Therefore for the deviation, caused by the metallic hull of ship, which are found on water, is correct expression (5.57).

When the direction-finding framework is establish/installed on the centerline of ship directly on deck, then m = 1. In other cases a radius of ship ρ_0 one should express by width and height/altitude of ship. Let us designate the width of ship in the site of installation of the direction finder by B and overall height of ship (from the upper deck to keel) by H. Approximately it is possible to accept for ρ_0 value

$$\rho_0 = \frac{B+H}{4}.$$

For the calculation of deviation, it is necessary to know the height/altitude of the installation of the framework of the direction finder above the upper deck. Let us designate it by h. Then for m one should write

$$m = \frac{\rho_0}{\rho} = \frac{\rho_0}{\rho_0 + h}. \qquad (5.60)$$

From formulas (5.58) and (5.60) follows that for decrease in the deviation, produced by the hull of ship (aircraft), it is expedient to rise the framework of direction finder as possible above above the hull of ship (aircraft).

Deviation from the hull of ship (aircraft) changes with a change in the wave. Only for the average/mean and long waves, which exceed the length of housing 10-15 times, deviation can be considered independent of wavelength and it it is possible to compute according to formula (5.58). For shorter waves the deviation increases at first, then it decreases.

The examined case of diffraction from semicylinder under terrestrial conditions approximately corresponds to the installation of radio direction finder at the flat/plane apex/vertex of the elongated elevation.

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5.8. Deviation of ship and aircraft radio direction finder.

As this follows from §5.5-5.7, the metal hull of ship (aircraft), and also any installations (antenna, masts, bridges, tubes, the locked ducts from delays with masts, etc.) aboard ship (aircraft) creates the fields of reradiation, the calling errors and the diffuseness of bearing.

In the constant/invariable position of object/subjects around external equipment/device of direction finder, the deviation also remains constant/invariable and it they consider with the use of radio direction finder. Deviation depends on wavelength.

Radio beam deviation f determines experimentally for the radio-course angles of q from 0° to 360° and depict in the form of curve. The curve of radio beam deviation can be expanded in Fourier series, is analogous with curved compass error:

$$f = A + B\sin q + C\cos q + D\sin 2q + E\cos 2q + \\ + K\sin 4q + L\cos 4q + \dots,$$

where A - the coefficient deviation constant

$$A = \frac{1}{2\pi}\int_0^{2\pi} f\,dq,$$

B and C - the coefficients of the semicircular deviation

$$B = \frac{1}{\pi}\int_0^{\pi} f\sin q\,dq \text{ и } C = \frac{1}{\pi}\int_0^{\pi} f\cos q\,dq,$$

D and E - the coefficients of one fourth deviation

$$D = \frac{1}{\pi}\int_0^{\pi} f\sin 2q\,dq \text{ и } E = \frac{1}{\pi}\int_0^{\pi} f\cos 2q\,dq,$$

K and L - the coefficients of octant deviation

$$K = \frac{1}{\pi}\int_0^{\pi} f\sin 4q\,dq \text{ и } L = \frac{1}{\pi}\int_0^{\pi} f\cos 4q\,dq.$$

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For explaining the reasons, calling the appearance of different coefficients of deviation, we proceed of the action examined above of separate emitters (antenna, the framework) on direction finding.

One should distinguish installation aboard the ship (aircraft) of the radio direction finder of medium-frequency waves and radio direction finder of short and ultra short waves.

On medium-frequency waves the deviation bears predominantly one fourth character. It is caused mainly by metal housing and it is little affected with a change in the wave. For a decrease in the deviation, it is necessary the antenna system of radio direction finder to rise above the hull of ship.

On short and ultra short waves curved deviations do not have this regular character as on medium-frequency waves. Greatest effect have antenna-like installations, especially with the size/dimensions, close to $1/4\lambda$ and $3/4\lambda$. Law and the value of deviation due to them depend on wave; therefore law and the value of deviation change with wave. For judging about deviation on these waves, one should proceed from the account of the closest antenna-like installations (masts, tubes, bridges, etc.) (see the curves of Fig. 5.14); it is necessary as far as possible to drive out from them the framework of radio direction finder up to this distance so as to bring deviation to the permissible value. This is achieved in the best way by arrangement/permutation by the antenna of the system of radio direction finder on the top of the mast of ship.

On medium-frequency waves usually it is possible to indicate the reasons for the appearance of different coefficients of deviation.

1. A, coefficient deviation constant,, does not depend on direction of incoming signal. It appears, if the indicator of the reading of bearing is establish/installed erroneously or in the presence of the ducts whose planes straddle of the framework of direction finder. If A is caused by last/latter reason, then its values change with a change in the wave. This case aboard ship (aircraft) corresponds to the installation of the framework not along the electrical axis of symmetry. In this case, the framework can be establish/installed from the centerline of ship, since the centerline of the symmetry of ship not always does coincide with its electrical axis of symmetry.

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2. B - coefficient of deviation, which is changed with period 2π, i.e., coefficient of semicircular deviation. It is caused by antenna-like emitter, arrange/located from the nose of ship (or aircraft) (+B) or from stern (-B).

3. C - coefficient of semicircular deviation. It is caused by antenna-like return emitters, which are located from the starboard (-C) or from the left side (+C) of ship or aircraft.

4. D - coefficient of quadrantal deviation. It is caused by the hull of ship (metallic aircraft, dirigible) or by the longitudinal and transverse ducts. Signs D are shown in Table 5.1.

5. E - coefficient of quadrantal deviation. Is caused by the ducts, arrange/located at an angle of 45° ~~ili~~ or 135° to the longitudinal axis. Signs E are shown in Table 5.1.

6. G and F - coefficients of sectant deviation, usually have very low value.

7. K and L - coefficients of octant deviation, depend on D and E.

As shown earlier, on the short and ultra short waves where affect the phase differences, caused by the distance between the directional antenna of radio direction finder and return emitters, the law of change f from q becomes complicated and usually it is not possible, on the basis of curved deviation, to indicate the reasons due to which appears the deviation.